

**SINGLE SIDEBAND, SUPPRESSED CARRIER  
RADIO COMMUNICATION**

**BY  
S. NEMAN**

**Thesis  
N37**

Library  
U. S. Naval Postgraduate School  
Annapolis, Md.









SINGLE-SIDEBAND, SUPPRESSED CARRIER  
RADIO COMMUNICATION

S. Neman





SINGLE SIDEBAND, SUPPRESSED CARRIER  
RADIO COMMUNICATION

by

S. Neman

Lieutenant Commander, United States Navy

Submitted in partial fulfillment  
of the requirements  
for a degree of  
MASTER OF SCIENCE  
in  
Electrical Engineering

United State Naval Postgraduate School  
Annapolis, Maryland  
1950

Thesis  
N44

This work is accepted as fulfilling  
the thesis requirements for the degree of

MASTER OF SCIENCE  
IN  
ELECTRICAL ENGINEERING

From the

United States Naval postgraduate School



## PREFACE

In 1923, the first radio communications link employing modulation in which the carrier and one sideband had been eliminated was established. Since then, the use of this system of transmission has grown to a considerable extent. The basic advantages of this system of communications are great. The major problems of the system have been overcome, although some problems do remain. And, too, there are some definite limitations to the practical applications of the system.

Because the writer believes this system holds much promise in the near future, this study of the nature and practical aspects of single sideband, suppressed carrier communications has been prepared.

The faculty of the Electrical Engineering Department of the United States Naval Postgraduate School has been extremely helpful in the preparation of this paper, and to them, the writer wishes to express his most sincere appreciation. In particular, he desires to thank Professor C. V. O. Terwilliger, chairman of the department, and Professor W. C. Smith, without whose guidance, assistance and encouragement this paper would have been impossible to prepare.



## TABLE OF CONTENTS

	Page
Preface	ii
List of Illustrations	iv
Chapter I    What is SSSC Communications?	1
Chapter II    SSSC Transmission	11
Chapter III   SSSC and SSRC Reception	27
Chapter IV    Advantages, Disadvantages, and Limitations of SSSC and SSRC	48
Bibliography	58
Appendix I	
Appendix II	
Appendix III	
Appendix IV	
Appendix V	
Appendix VI	





## LIST OF ILLUSTRATIONS

No.	Page
1(A) Block Diagram of A.M. Modulator	4
1(B) Single Modulating Frequency Representation of A.M.	4
1(C) Representation of Spectrum of A.M. with Finite Audio Range	4
2 Character of SSSC Wave	7
3 Process of Production of SSSC Wave	12
4 The Balanced Modulator	13
5 The Van der Bijl Modulator	13
6 Typical Triode Characteristics	13
7(A) The Ring Modulator	18
7(B) Equivalent Circuit of Ring Modulator for Conduction in one Direction	18
7(C) Equivalent Circuit of Ring Modulator for Conduction in other Direction	18
8 Required Attenuation Characteristics for a Typical Sideband Filter	21
9 Vector Diagram Analysis of Phasing Method of SSSC Production	23
10 Block Diagram of Phasing Method of SSSC Production and Corresponding Vector Diagrams	26
11 Balanced Frequency Converter for SSSC Reception	37
12 A Block Diagram of a Single-Sideband Receiving System	40



	Page
13 Separation of Sidebands by Phasing Method -- Upper and Lower Sidebands	41
14 Effect of Audio Filter on Selectivity Characteristics of a Receiver	47
15 Block Diagram of a Conventional A.M. Transmission System	48
16 Block Diagram of SSSC Transmission System	50



## TABLE OF SYMBOLS AND ABBREVIATIONS

These are listed in the order of their use in the text.

A.M.	-- Amplitude Modulated
F.M.	-- Frequency Modulated
P.M.	-- Phase Modulated
KC	-- Kilocycles
$e$	-- Instantaneous Voltage
$M$	-- Modulation Factor (Degree of Modulation)
$\omega$	-- $2\pi$ Times Frequency in Cycles per Second
$\omega_s$	-- $2\pi$ Times Signal Frequency
$\omega_c$	-- $2\pi$ Times Carrier Frequency
$e_c$	-- Instantaneous Carrier Voltage
$e_s$	-- Instantaneous Signal Voltage
$e_o$	-- Instantaneous Resultant Voltage
SSSC	-- Single-Sideband, Suppressed Carrier
SSWC	-- Single-Sideband, With Carrier
SSRC	-- Single-Sideband, Reduced Carrier
R.F.	-- Radio Frequency
$i_p$	-- Instantaneous Plate Current
$R$	-- Resistance
$i$	-- Instantaneous Current
$e_g$	-- Instantaneous Grid Voltage



$e_p$  -- Instantaneous Plate Voltage

CW -- Carrier Wave (Code)

BFO -- Beat-Frequency Oscillator

C.P.S. — Cycles per Second

KW — Kilowatts





## CHAPTER I

### WHAT IS SINGLE SIDEBAND SUPPRESSED CARRIER COMMUNICATIONS?

This paper concerns itself with a type of modulation of a radio-frequency carrier wave. Modulation, as is very well known, is the process of introducing intelligence into a fixed frequency, constant amplitude carrier wave. Several methods for accomplishing this end are in general use, each having it's advantages and limitations. A brief discussion of the most common modulation methods is in order at this time.

Three methods of modulation are widely used in radio communications. These are amplitude modulation (A.M.), frequency modulation (F.M.), and phase modulation (P.M.). The first, A. M., is the most commonly used method. All radio broadcasting stations in the 550 KC to 1700 KC band use A.M. Most communications traffic also uses this modulation method. In principle and practice, this method of modulation is most simple, thus accounting for the widespread use. A.M. is usually obtained by superimposing the desired audio signal on the plate power supply of a non-linear, radio-frequency amplifier. The resultant output from this stage is a signal of variable amplitude, which may be represented by the equation,

$$e = A(1 + M \cos \omega_s t) \cos \omega_c t \quad (\text{Eq. I})$$



in which M is the modulation factor (degree of modulation, from 0 to 1),  $\omega_s$  is the signal, or audio frequency, and  $\omega_c$  is the carrier frequency. Equation I is derived as follows:

$$e_c = A \cos \omega_c t = \text{carrier}$$

$$e_s = B \cos \omega_s t = \text{modulating signal}$$

$$e_o = (1 + e_s)e_c = A(1 + B \cos \omega_s t) \cos \omega_c t.$$

It will be noticed that the degree, or percent, of modulation is dependent on the value of B. Maximum, or 100% modulation will occur when the minimum amplitude of the modulated wave is zero. (A greater value of B will result in "over-modulation" which produces distortion and spurious radiation. It is not acceptable.) This value will occur when B = 1, so that B may have a value from 0 to 1. M, the percent modulation, is defined as equal to B, thus

$$e_o = A(1 + M \cos \omega_s t) \cos \omega_c t \quad (\text{Eq. I})$$

Expanding this equation,

$$\begin{aligned} e_o = A \cos \omega_c t + 1/2 AM \cos (\omega_c + \omega_s)t \\ + 1/2 AM \cos (\omega_c - \omega_s)t. \end{aligned} \quad (\text{Eq. II})$$



Thus it is seen that an A. M. transmission contains components of three frequencies, the carrier, and the sum and difference of the carrier and signal frequencies. These sum and difference frequencies are known respectively, as upper and lower sidebands. Figure 1 may assist in visualizing A. M.

Frequency modulation is the next most popular method of modulation. In the last decade, it has become very popular because of certain inherent favorable characteristics. A widely used method of obtaining F. M. is the use of a "reactance tube" shunting the tank circuit of an oscillator. The output may be represented by

$$e_o = A \sin \left( \omega_c t + \frac{2\pi \Delta f}{\omega_s} \sin \omega_s t \right)$$

Where  $\Delta f$  is the maximum deviation of the instantaneous frequency from the carrier frequency and is determined by the peak magnitude of the modulating signal. It is independent of the modulating frequency. The expansion of this equation, while at first glance may appear to contain components of only a limited number of frequencies, actually contains an infinite number of frequencies, as is seen in the expanded equation below.



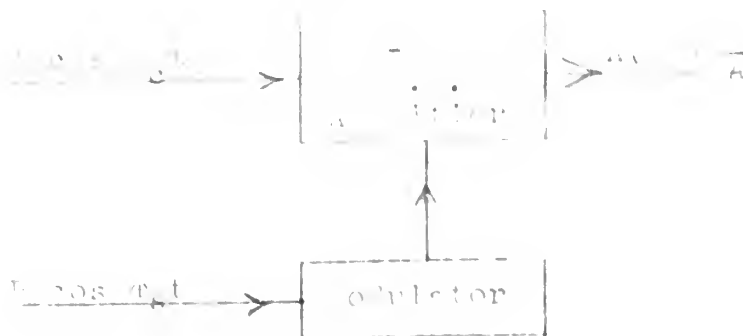


FIGURE 1(A). Block diagram of modulator



FIGURE 1(B). Single modulation frequency representation of A. W.

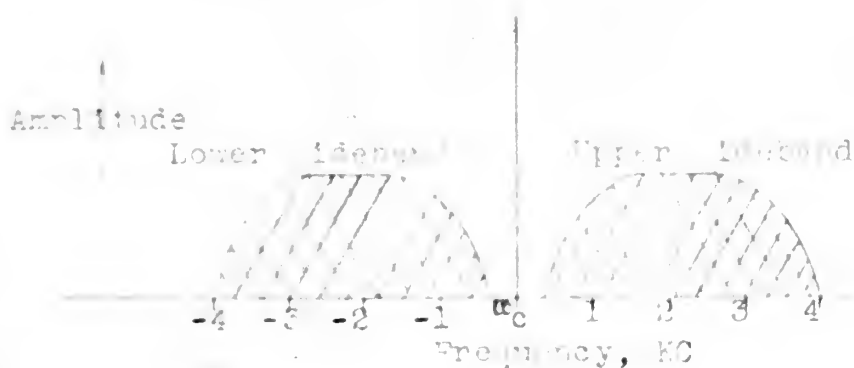


FIGURE 1(C). Representation of Spectrum of A. W. with Finite Audio Range





$$\begin{aligned}
e_o = A \Big\{ & J_0(m_f) \sin \omega_c t \\
& + J_1(m_f) \left[ \sin(\omega_c + \omega_s)t - \sin(\omega_c - \omega_s)t \right] \\
& + J_2(m_f) \left[ \sin(\omega_c + 2\omega_s)t + \sin(\omega_c - 2\omega_s)t \right] \\
& + J_3(m_f) \left[ \sin(\omega_c + 3\omega_s)t - \sin(\omega_c - 3\omega_s)t \right] \\
& + \dots \dots \dots \Big\}
\end{aligned}$$

where  $m_f = 2\pi A f / \omega_s$  = modulating index, and  $J_n(m_f)$  means the Bessel function of the first kind and nth order, with argument  $m_f$ . Thus it is seen that the frequency spread is wide, indeed. In actual practice, however, the frequency spread is approximately  $(\omega_s + 4f)$  on each side of the carrier frequency, with components spaced at frequency intervals that are equal to the modulating frequency.

Phase modulation is similar to frequency modulation in many respects. The equations of  $e_o$  are the same for phase modulation as for frequency modulation. The only difference is in the meaning of  $m$ , which for phase modulation should have a subscript of  $p$ .  $M_p$  is the angle in radians through which the phase is displaced, at the peak of the modulation cycle, away from the phase that would exist if there were no modulation.

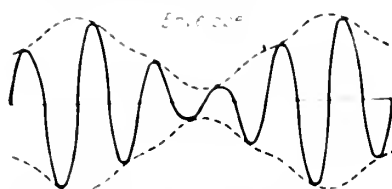


This paper is not concerned with frequency or phase modulation, and no further mention will be made of them. Returning to Eq. II, it is again noted that  $e_0$  consists of three components. The carrier, of course, contains no intelligence. Each of the sidebands, however, contains all of the intelligence. It would seem logical, then, that if only the sidebands, or even only one sideband, were transmitted, all the intelligence would be contained. The immediate value of this type of transmission would be to greatly reduce the width of the transmission band for a given range of intelligence frequencies. This has many advantages over the other types of transmissions which, for certain types of communications, make it well worth the practical difficulties encountered in the production and reception of this type of transmission. This is Single-Sideband, Suppressed Carrier Communications (SSSC). Visualization of the waveform of SSSC is somewhat difficult, but Figure 2, which illustrates this with an excessively low carrier, may assist. Under certain conditions, it is desirable to transmit all or part of the carrier. This mode of operation will be called Single Sideband With Carrier (SSWC), or Single Sideband, Reduced Carrier (SSRC). A third possibility is the transmission of A.M. with only the carrier suppressed. Transmission in this manner is possible, and is in use for some applications, but involves additional reception problems which will be discussed later.

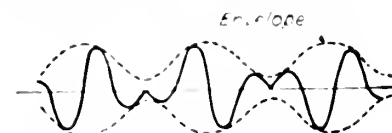




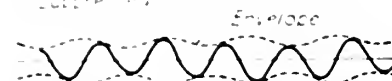
(a) Original signal consisting of fundamental and its first harmonic



(b) Original signal modulated on carrier wave



(c) Modulated wave after carrier has been suppressed



(d) Modulated wave after suppression of carrier and one sideband

**FIGURE 2.1** Character of waves produced when carrier is suppressed and when carrier and one side band are both suppressed. For purposes of clarity, the carrier frequency is made excessively low.



This paper will be a discussion of the nature of, and practical problems encountered in the transmission and reception of SSSC, SSWC, and SSRC. Before going into the details of SSSC operation, the necessary enthusiasm may be aroused by a quick look at some of the advantages. If a pure resistive load is assumed for the transmitter, the power output of the transmitter is  $E_0^2 R$ . For an A.M. signal,

$$e_0 = A \cos \omega_c t + AM/2 \cos (\omega_c + \omega_s)t + AM/2 \cos (\omega_c - \omega_s)t$$

$$E_0^2 = A^2 + (AM/2)^2 + (AM/2)^2 .$$

Thus the power output,

$$P_0 = [1 + (M/2)^2 + (M/2)^2] RA^2 .$$

The power distribution among the three components of the A.M. output is:

$$\text{Carrier} = (1 - \frac{M^2}{2})100\% / (1 + \frac{M^2}{2})$$

$$\text{Upper Sideband} = (M^2/4)100\% / (1 + \frac{M^2}{2})$$

$$\text{Lower Sideband} = (M^2/4)100\% / (1 + \frac{M^2}{2})$$

For two values of M, the power distribution is:

	M = 1	M = .25
Carrier	66.7%	96.8%
Each Sideband	16.7%	1.56%

As mentioned earlier, overmodulation produces spurious frequency emissions. This is not only undesirable, but





is also illegal. To reduce the possibility of overmodulation on peaks of audio power, the percent modulation is normally kept at a much lower value than 100%. A value of 25% is normal. But at 25% modulation, the percent useful power is only 1.56%! This is only 780 watts useful power out of 50,000 watts transmitted power.! As one author<sup>13</sup> aptly put it, of this 50,000 watts of steam being generated, 48,400 watts is being used to toot the whistle!

The greatest improvement in this situation is realized by elimination of the carrier. Immediately, the percent power in each sideband becomes 50%. This is a feasible method of operation and is used to a great degree in wired communication systems. As will be explained in the chapter under Reception, this method of transmission introduces considerable difficulty in reception. Elimination of one sideband in addition to the carrier not only raises the percent usable power to 100%, but greatly reduces the reception problems introduced by carrier suppression. An advantage of even greater importance in many applications is the reduction of required frequency by half. The available frequency spectrum is not unlimited. With the constantly increasing demands for transmission space, some method of doubling the available space is not to be viewed lightly. For wired communication circuits, the spectrum is also limited by the physical constants of the transmission line.



The narrower the required band, the larger will be the number of messages per circuit. This method of modulation is most widely used by telephone companies, as would be expected.

Another not-too-obvious advantage of this system is the savings in transmitter cost. Most high-powered transmitters are modulated by plate modulating the final power amplifier. To accomplish this, an audio system is required that delivers half the input power to the final power amplifier. For a 50,000 watt transmitter, this means an audio power of approximately 25,000 watts is required. Audio equipment in this power classification will be a major item in the overall cost of the transmitter. When SSSC is employed, modulation takes place in the lowest powered stages. Receiving-type equipment may be used for all the audio equipment, thus greatly reducing the equipment cost.

There are other advantages in the use of SSSC, and they will be discussed later.



## CHAPTER II

### SSSC TRANSMISSION

The problem of generation of SSSC resolves itself into separate stages. First is the elimination of the carrier. Next is the suppression of one of the sidebands. In some circuits, the order of these stages is reversed or even split, but the two stages are always distinct. This process may be represented by the block diagram of Figure 3.

The oscillator, or oscillators (as it will be seen later, several oscillators may be required) will not be discussed at this time except to mention that they must be of very stable design.

The modulator and carrier suppressor is so listed, because the use of a single unit of proper type will accomplish both requirements. Several types of units are used for this purpose. The first, and most commonly used, is the balanced modulator. Figure 4 is a simplified schematic diagram of a balanced modulator. Consideration of this circuit, however, will be made easier by first considering Figure 5, the Van der Bijl modulator.

In the Van der Bijl modulator, as in the balanced modulator, the circuit is that of a Class A, R. F. amplifier with the audio signal input fed into the stage in series with the carrier frequency input. It will be



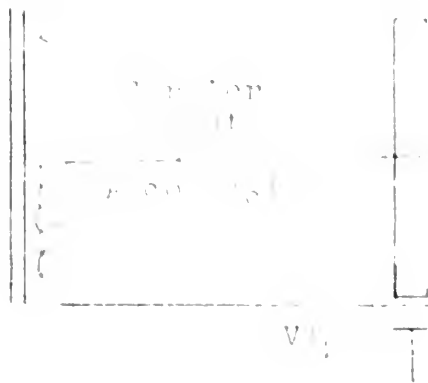


FIGURE 3. PROCESS OF PRODUCING OF 5510 A.E.





Signal  
Voltage  
Input  
 $E \cos \omega_s t$



Carrier  
Input  
 $E \cos(\omega_c - \omega_s)t +$   
 $K \cos(\omega_c + \omega_s)t$

FIGURE 4. Carrier Input

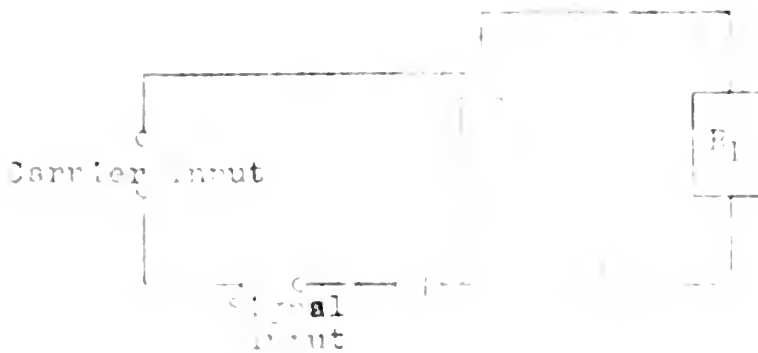


FIGURE 5. Carrier Input

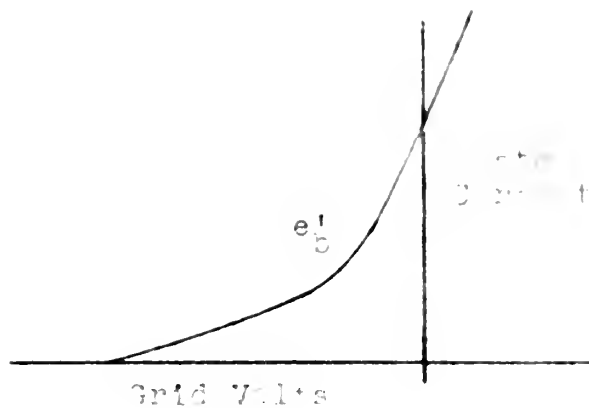


FIGURE 6. Typical Grid Characteristics



remembered that the characteristics of a triode are non-linear, as indicated by Figure 6. This curve may be expressed as a power series. Considering only the first two terms of this power series,

$$i_p = a_1 e + a_2 e^2 = \text{plate, or load current}$$

where

$$e = e_c + e_s = B \cos \omega_s t + A \cos \omega_c t = \text{exciting voltage.}$$

Then:

$$i_p = a_1 (A \cos \omega_c t + B \cos \omega_s t) + a_2 (A \cos \omega_c t + B \cos \omega_s t)^2.$$

By expansion and trigonometric rearrangement,

$$i_p = a_1 A (1 + \frac{2a_2}{a_1} B \cos \omega_s t) \cos \omega_c t + a_1 B \cos \omega_s t + a_2 B^2 \cos^2 \omega_s t + a_2 A^2 \cos^2 \omega_c t.$$

Now note that the first term of the last equation is identical in form to Eq. I, which is the equation of an amplitude modulated wave. The remaining three components are removed from the circuit by suitable filtering.



Returning to the balanced modulator of Figure 4, it is seen to be a straightforward push-pull stage of Class A amplification to which a source of carrier voltage has been added. Considering the audio source only, it is seen that it is amplified in a conventional manner and will appear across the primary of  $T_2$ . The carrier voltage, however, is impressed on the grids of  $VT_1$  and  $VT_2$  in the same phase, producing currents, as indicated in Figure 4, in the same direction with respect to the plates of the tubes. These currents induce voltages in  $T_2$  that ~~cancel~~, thus the carrier does not appear in the output. The sidebands,  $kAB \cos(\omega_c + \omega_s)t$  and  $kAB \cos(\omega_c - \omega_s)t$  will appear, however. This may be shown mathematically as follows:

$$\text{Exciting voltage of } VT_1 = e_c + e_s$$

$$\text{Exciting voltage of } VT_2 = e_c - e_s$$

$$i_{p1} = a_1(e_c + e_s) + a_2(e_c + e_s)^2 + a_3(e_c + e_s)^3 \dots\dots\dots$$

$$i_{p2} = a_1(e_c - e_s) + a_2(e_c - e_s)^2 + a_3(e_c - e_s)^3 \dots\dots\dots$$

Note, now, that  $i_{p1}$  and  $i_{p2}$  flow in the primary of  $T_2$  in opposite directions, so that the voltage induced in the secondary is proportional to  $i_{p1} - i_{p2}$ .



$$\begin{aligned}
e_o &= K(i_{p1} - i_{p2}) \\
&= 2A \left[ a_1 e_s + 2a_2 e_c e_s + a_3 (3e_c^2 e_s + e_s^3) \right. \\
&\quad + a_4 (4e_c^3 e_s + 4e_c e_s^3) + a_5 (5e_c^4 e_s + 10e_c^2 e_s^3 + e_s^5) \\
&\quad \left. + \dots \dots \dots \right].
\end{aligned}$$

Now,  $a_3$ ,  $a_4$ , and  $a_5$  are small in comparison with  $a_1$  and  $a_2$ . Also, the third and higher order terms become smaller as  $e_s$ , or percent modulation, becomes smaller. The effect of all terms above the second order may thus be made negligible. The equation then reduces to

$$\begin{aligned}
e_o &= 2K(a_1 e_s + 2a_2 e_c e_s). \\
\text{As } e_c &= A \cos \omega_c t, \quad , \text{ and} \\
e_s &= B \cos \omega_s t, \\
e_o &= 2Ka_1 B \left( 1 + 2 \frac{a_2}{a_1} A \cos \omega_c t \right) \cos \omega_s t
\end{aligned}$$

which, by trigonometric rearrangements, gives

$$\begin{aligned}
e_o &= 2Ka_1 B \cos \omega_s t + 1/2 \left[ 4Ka_2 AB \cos (\omega_c + \omega_s)t \right. \\
&\quad \left. + 4Ka_2 AB \cos (\omega_c - \omega_s)t \right].
\end{aligned}$$

Note that this output contains only the sidebands plus the modulating signal frequency component. This latter component is attenuated sufficiently in  $T_2$  so as to be negligible.





In this derivation, perfect balance of tube and circuit elements has been assumed. In practice, it is usually necessary to provide adjustable balancing elements to the circuit, as well as to use care in selection of balanced tubes.

The functions of carrier suppression and modulation may also be accomplished by means of a circuit known as the "ring" modulator. As shown in Figure 7(A), this consists of four diode rectifiers, usually of the contact type, connected in a series ring. The nature of most diodes is such that with application of a forward voltage, the effective resistance  $R_f$ , will be small. With the application of a reverse voltage, the effective resistance,  $R_r$ , will be large. The magnitude of  $e_c$  is made very much greater than  $e_s$ , so that in effect,  $e_c$  is simply a switch that switches resistances between the positions in the equivalent circuits of Figure 7(B) and (C). It may readily be shown that efficiency is greatest for  $R_s = R_s = R_L = \sqrt{R_f R_r} = R$ . To simplify the following work, it will be assumed that this condition exists. Noting that  $i_a = e_s/2R$ , the solution of the network for  $i_b$  becomes:





-----



-----



3. In the case of a circuit in which the current is in one direction only, the direction of the current is the same as the direction of the magnetic field.



$$i_b = \frac{e_s}{2R} \frac{\sqrt{\frac{R_r}{R_f}} - 1}{\sqrt{\frac{R_r}{R_f}} + 1} f(t) ,$$

where  $f(t) = \begin{cases} +1 & \text{for positive half cycle of carrier} \\ -1 & \text{for negative half cycle of carrier.} \end{cases}$

Representing  $f(t)$  by Fourier components,

$$f(t) = \frac{4}{\pi} \left( \cos \omega_c t - \frac{1}{3} \cos 3\omega_c t + \frac{1}{5} \cos 5\omega_c t - \dots \right).$$

Substituting  $e_s = B \cos \omega_s t$ .

$$i_b = \frac{4B}{2R} \frac{\sqrt{\frac{R_r}{R_f}} - 1}{\sqrt{\frac{R_r}{R_f}} + 1} \cos \omega_s t \left( \cos \omega_c t - \frac{1}{3} \cos 3\omega_c t \dots \right).$$

Now, as  $B$  is very small, the use of simple circuit elements will reduce components over the first order to a negligible value, leaving

$$\begin{aligned} i_b &= K \cos \omega_s t \cos \omega_c t \\ &= \frac{K}{2} \cos(\omega_c + \omega_s)t + \frac{K}{2} \cos(\omega_c - \omega_s)t. \end{aligned}$$

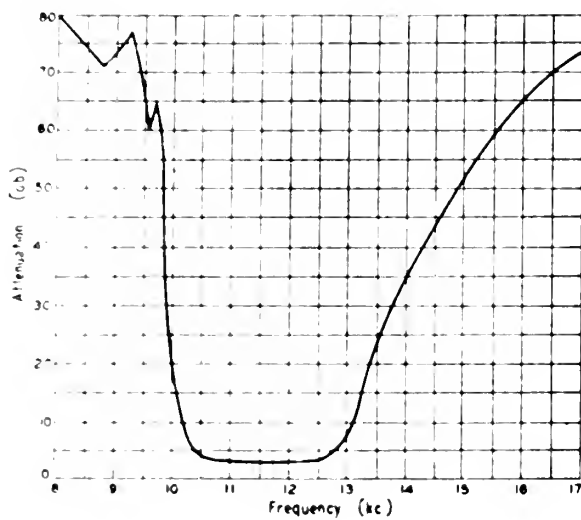
Thus, not only has the carrier been eliminated, but also so has the modulating signal.



The next stage of the transmitter is the sideband eliminator. Only two methods are in general use for this job. The most popular is the straight filter method. The sideband eliminator consists of a suitable filter to stop the unwanted sideband. This is known as ~~the~~ "brute force" type of sideband eliminator to contrast it with the more subtle, but also more critical and delicate, phasing method of sideband elimination. Figure 8 shows the necessary frequency characteristics of a suitable filter. The required sharpness is not a percent function of the carrier frequency, but is a function of the number of cycles per second. This would imply that the lower the carrier frequency, the less would be the required percent sharpness, and thus the simpler would be the filter design. In practice, if the initial carrier frequency is kept below about 40 KC, a conventional, multiple, M-derived band-pass filter is adequate. Figure 8 is the characteristic curve of such a filter using a 10KC initial carrier frequency. If the initial carrier frequency exceeds 40 KC, more complex filters of the crystal-lattice type are required. A ceiling of approximately 100 to 150 KC exists for practical filters. Discussed in this paper will be only filters for use with carrier frequencies up to 40KC. This low frequency carrier modulated by a balanced modulator, which eliminates the carrier, then one sideband







(National Company, Inc.)



is filtered out. Appendix I describes the design and construction of such a filter. The resulting SSSC wave is then boosted to final transmitter frequency in one or more stages by mixing, or heterodyning with one or more oscillators of suitably higher frequency, each followed by an adequate filter to remove the unwanted new sideband. This latter filter is very simple, consisting of little more than a tuned amplifier stage, or a simple tuned circuit. The filter method of sideband elimination, then in reality consists of two or more oscillators and balanced modulators (or mixers) in addition to the main and secondary filters.

The second method of sideband elimination, by phasing, may be more readily understood if the vector diagrams of Figure 9 are first studied. From the equation of an A.M. wave (100% modulation used for convenience), it is seen that the final voltage consists of the same three components previously mentioned. As these are of different frequencies, namely  $\omega_c$ ,  $(\omega_c + \omega_s)$ , and  $(\omega_c - \omega_s)$ , any single vector diagram representing all these components should indicate the rotating frequency, and it should be remembered that the vector diagram represents only the instantaneous position of the vectors -- that their relative position will change from instant to instant. Now, for any random instant of time, the vectors may have the relative position as shown in Figure ((A)).

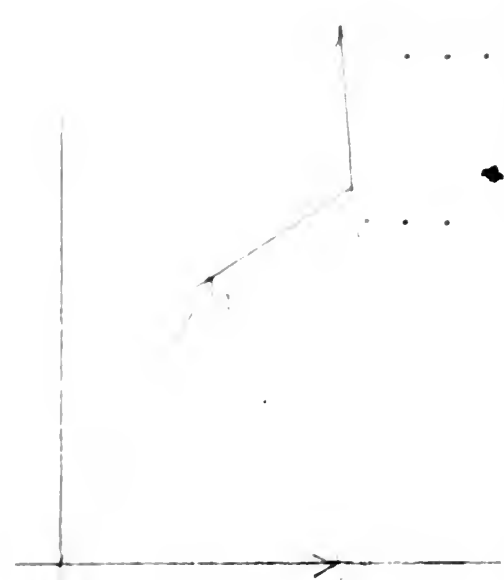




(11).



(12).



(13).



(14).



(15).



The same position may be constructed as shown in Figure 9(B), by noting that with the sideband vectors now originating from the moving point, the end of the carrier vector, their frequency of rotation about the moving point will be their frequency of rotation about the origin minus the frequency of rotation of the moving point. For the upper sideband, the frequency is  $(\omega_c + \omega_s) - \omega_c$ , or just  $\omega_s$ . For the lower sideband, the frequency is  $(\omega_c - \omega_s) - \omega_c = -\omega_s$ . To obtain the value of  $e_0$ , it is necessary to vectorally add the components, as is done in Figure 9(C).

For the position of peak value of  $e$ , the vector position is that of Figure 9(D). Let this be called the reference position. Suppose, now, that to a signal having the reference position, a signal, which at the same instant of time has a vector position as shown in Figure 9(E), is added. It is further assumed that the respective lengths of vectors are identical. On adding, it is seen that the lower sideband vectors cancel each other, while the upper sideband vectors add. The carrier components also add in quadrature, of course, but this is not significant. The important point is that one sideband has been eliminated. The problem, then, resolves itself into the production of two amplitude modulated signals,  $e_0$  and  $e_1$ , that have the indicated phase differences. This may be accomplished in two steps. First, take a signal equal to  $e_0$  and





shift it through 90 degrees, as indicated by Figure 10(A). Next, shift the audio input by 90 degrees, which results in Figure 10(B) and is the required wave of Figure 9(E). The block diagram of Figure 10(C) indicates a method of generating the SSSC output by this phasing method. The use of balanced modulators eliminates the carrier. Of course, all components must be accurately balanced for satisfactory operation. The audio phase shifter is the only item not of routine design. Appendix II is a published article<sup>4</sup> on this subject and presents a basis for design.





Figure 10(1).

Figure 10(2).

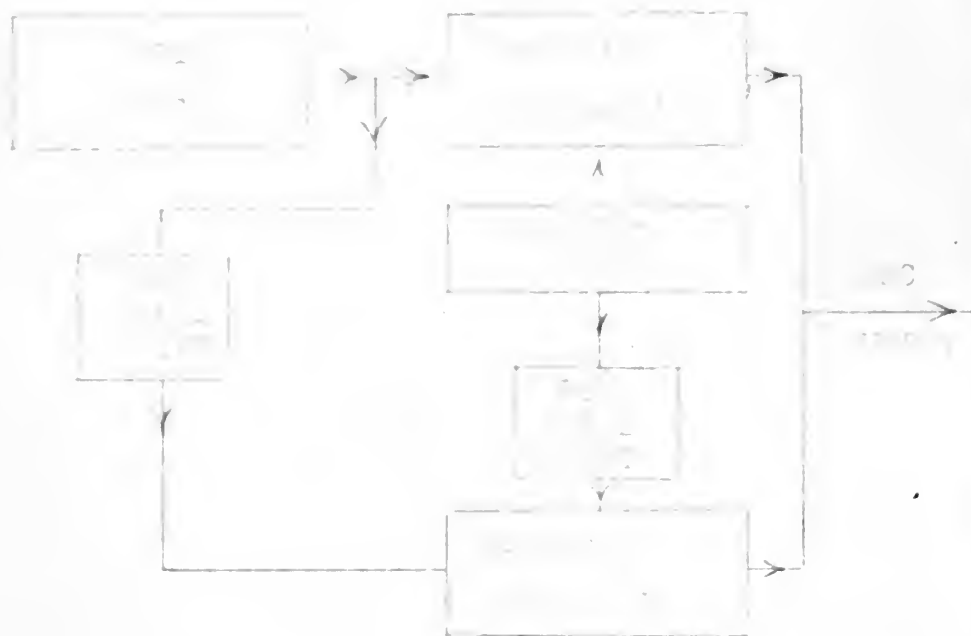


Figure 10(3).



## CHAPTER III

### SSSC AND SSRC RECEPTION

Before considering reception of SSSC or SSRC, the subject of detection, or demodulation, of modulated waves in general should be discussed. Detection, as principally used, is of three types -- square-law, linear, and hetrodyne. The first two of these are closely related.

Referring again to Figure 6, it is seen that a typical characteristic curve of a triode ( $i_p$  vs.  $e_g$ ) is made up of essentially straight section, followed by a knee, followed by another more-or-less straight section. The shape of this curve is also representative of the  $i_p$  vs.  $e_p$  characteristic curve of a diode. If a detector is operated principally in the region of the knee of the characteristic curve, it is called a square-law detector. The name is derived from the fact that this portion of the curve may be closely approximated by a second order equation. If, on the other hand, the voltage variations are very large, then the effect is to shrink the scale of Figure 6, so that the region of the knee shrinks in effect to a point. Thus Figure 6 could be represented by two intersecting straight lines. This type of detection is called linear



detection. Because linear detection necessarily involves large values of signal-voltages, the term "power detection" is sometimes applied. Conversely, the small values of signal-voltage used in square-law detection leads to the name of "weak-signal" detector. An arbitrary dividing point of about one volt is generally made.

In practice, a square-law detector usually employs a tube containing a control grid in order to obtain amplification in the stage as well as detection. The circuit of such a detector would be that of an amplifier except that voltages impressed are such as to cause the tube to operate principally in the region of the knee of the characteristic curve. Mathematical analysis of the detection process is rather simple. If  $e_o$  is the impressed modulated wave, and  $i_p$  is the detector plate current, for square-law detection,

$$i_p = a_1 e_o + a_2 e_o^2 .$$

For an A.M. wave,

$$e_o = E_c \cos \omega_c t + 1/2 E_c M \cos(\omega_c + \omega_s)t \\ + 1/2 E_c M \cos(\omega_c - \omega_s)t.$$





Case I of Appendix III carries out the derivation of the output components. The balance of Appendix III repeats the procedure for different conditions of  $e_0$ . A summary of the results of Appendix III appear in Table I.

As was mentioned earlier, the results appearing in Table I indicate that reception of A.M. with only the carrier suppressed would be difficult in that, as shown in Cases II, III, and IV, the carrier must be reinserted at exactly the same frequency, and closely in phase with the original carrier, if the resulting signal is to be intelligible. Except with the most expensive equipment, this is difficult, or impossible to accomplish.

Mathematical analysis of linear detection is not so simple or straightforward as was that for square-law detection. Because of the discontinuity in the effective characteristic curves for linear detection, a Taylor power series may not be used to represent the entire curve. A convenient mathematical dodge may be used by considering only the audio components. These audio components will be contained in the equations of the upper and lower envelopes of the modulated wave. Appendix IV carries this method of analysis out first for a square-law detector to demonstrate the method, then for linear detection of an A.M. wave. Linear detection (and square-law detection also) of a SSSC wave in reality is the



I	II	III	IV	V
I	...	---	...	...
II	...	...	...	intelligible
III	...	...	...	Excellent if $\Delta$ is small. Otherwise, reduce $\Delta$ to zero if $\Delta$ is large.
IV	...	---	...	intelligible
V	...	---	---	...
VI	...	...	...	...
VII	...	...	...	Excellent for $\Delta < 10$ Fair for $\Delta < 20$

$$K_1 = \frac{1}{2} \frac{L}{C} : K_2 = \frac{1}{2} \frac{L}{C} : K_3 = \frac{1}{2} \frac{L}{C}$$

Figure 1. Average values (estimated from Figure 1.1)



third type of demodulation -- hetrodyning -- and will be discussed in the next section.

Comparing the results of detection of an A.M. wave by square-law and linear detection, two conclusions of importance are apparent. First, square-law detection generates a second harmonic of the signal, which distorts the output signal to a degree equal to  $M/4$ , while no such component is produced by linear detection. Next, the magnitude of the signal frequency output (audio) of a square-law detector is proportional to the square of the carrier magnitude, while for a linear detector, it is proportional to the first power. These conclusions have resulted in the almost universal adoption of linear detection for conventional A.M. receivers because less distortion results, and the receiver is less affected by signal fading conditions. This situation may appear irrevelant at this time, but if SSSC reception is to be attempted with a conventional receiver, it becomes important, as will be shown later.

Hetrodyning, the third method mentioned of demodulation, is the process of combining a locally generated wave with the incoming wave in such a manner that a third wave is produced which either is the signal component, or contains the signal component in modulated form. This device is used



in superhetrodyne receivers to convert the incoming wave to a lower (fixed and predetermined) frequency. It is also used in communications receivers to supply an audible signal when CW (code) is being received. Appendix V, taken from "Theory and Applications of Electron Tubes" by Reich, is an excellent discussion of the nature and process of this type of detection. As indicated there, the simple mixing of two waves of different frequencies would result in a beat-frequency wave which had much of the appearance of a conventional A.M. wave, however, it is a wave that is varying in phase as well as in magnitude. The envelope, which is of the difference frequency, can be obtained in the form of useful power only by sending the two waves through a non-linear circuit element, such as a square-law or linear detector. (As has been implied, the required non-linearity in a linear detector is provided by the discontinuity in the characteristic curve.) This process, in effect, amounts to the process of modulation, not demodulation. Sidebands of the sum and difference frequencies are produced which is the same as the result of A.M.

It is apparent now, that demodulation of SSSC will always be of the hetrodyning type, regardless of whether





a linear or a square-law detector, or mixer, is used. The process is basically one of mixing the incoming wave with a locally generated wave to give the audio frequency output. For detection using a square-law mixer, Cases VI and VII of Appendix III present a derivation of the resulting output components. The difference frequency is the audio frequency signal. No second harmonic distortion signal is present. Referring again to Appendix IV, it is seen that for a linear mixer, the output components are infinite in number, containing an infinite number of audio distortion components. The amplitudes of these distortion components are functions of  $h^{n-1}$ , where  $h$  is the ratio of the amplitudes of the incoming signal to the locally generated signal, and  $n$  is the order of the harmonic. The resultant amount of distortion becomes acceptable at values of  $h$  below 2. The type of harmonic distortion described here will be referred to as "rectification distortion".

We are now able to consider practical reception of SSSC. A conventional A.M. communications receiver is equipped with a local oscillator (in addition to the one used in the superhetrodyne circuit) to be used in reception of CW signals, as was mentioned above. This is usually provided with an external frequency adjustment control, so that it



amounts to a local, variable frequency, beat-frequency oscillator (BFO). Its frequency range is from somewhat above, to somewhat below the incoming carrier frequency after conversion to the intermediate frequency in the super-hetrodyne mixer circuit. Thus, this oscillator should be usable in the demodulation of SSSC. When a SSSC wave is tuned in, only a garbled gibberish will be heard. This signal must be carefully tuned to maximum strength with the BFO and the automatic volume control switched off. Now the BFO is switched on -- the tuning control is not touched again. The frequency adjustment control of the BFO is slowly adjusted until the gibberish becomes intelligible. If this does not occur, the gain of the R.F. stages should be reduced to as low a value as possible. In fact, normally, the volume control should be on maximum, and the R.F. gain adjusted to just enough gain to provide adequate volume. The reason for this is that practically all conventional receivers employ linear detection. When this linear detector is used as a linear mixer for SSSC demodulation, then as has already been shown, the value of  $h$  (ratio of amplitudes of incoming signal and local oscillator) must be below .2 if the resultant distortion is to be acceptable. A lower value would be still more



desirable. If, after reducing the R.F. gain to as low as value as possible, adjustment of the BFO control does not result in an intelligible signal, then either the BFO control does not permit a wide enough frequency variation, or the amplitude of the BFO is too small to reduce  $h$  to an acceptable value. If such is the case, then the receiver must be modified before SSSC transmissions may be received.

As indicated in Table I, the frequency of the local oscillator must be within less than 50 c.p.s. of the original carrier. It is desirable for it to be within less than 10 C.P.S. If music were being transmitted, the frequency difference would have to be less than 5 c.p.s. Thus it should be obvious that the local oscillator must have very good stability to prevent fading in and out of intelligibility. This degree of stability is provided in only the best of communications receivers, as it is not required for A.M. reception.

From what has been said above, it seems that the use of a conventional receiver for reception of SSSC, while possible with most such receivers, is at best, only a makeshift arrangement. This type of receiver is poor for two reasons -- the type of detector and the characteristics of the BFO. With a good quality communications receiver,



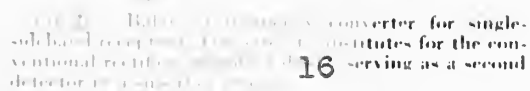
satisfactory SSSC reception is usually possible with only the addition of an external oscillator of very stable design and of sufficient amplitude.

In some cases, however, the high amplitude of the auxiliary oscillator overloads the detector, and thus introduces additional distortion. If, in addition, the detector is changed to a more suitable type, then reception is possible that, while not as good as that of the huge receivers employing multiple crystal filters and fixed frequency operation, is very good indeed. An excellent detector for this use is shown in Figure 11. Comparing Figure 11 with Figure 4 indicates that the demodulator is a form of the balanced modulator. As even for square-law mixers, some rectification distortion will occur, the use of a balanced mixer of the square-law type permits the cancelling out of the second harmonic distortion components, thus the advantages of both the linear and the square-law mixers are obtained with none of the disadvantages.

The problem of design of a local oscillator that is stable to within several c.p.s. is a difficult one. In addition, the additional complication of the tuning process due to having to adjust the frequency of the local oscillator for each station received can be serious if







( 4 )

(1) (a)  $\{1, 2, 3, 4, 5, 6, 7, 8, 9, 10\}$

18 — (Continued)

11. — 1,0000 column

$$\mathbf{B}_3 = \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{pmatrix} \quad \text{and} \quad \mathbf{B}_4 = \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{pmatrix}.$$
 $R_0, R_1 = 10,000 \text{ dollars}$ 

1-1-1 transformer with center-tapped secondary coupled to beat oscillator.

1 • Audio transformer—push-pull **primary** capable of carrying about 10 ma.



rapid change of station is necessary. If the carrier were only reduced, say 20 db, rather than eliminated, then the carrier would be received. Practically all the advantages of SSSC transmission would still be realized, and the small amount of carrier received could be used to provide a synchronizing signal for the local oscillator. The advantages are obvious. The disadvantage is that reception of this transmission on a receiver not having a synchronizing device would be disturbed by a low-frequency beat signal formed between the carrier transmitted and the reinserted carrier. This disturbance would be of relatively small magnitude if the transmitted carrier were reduced as much as 20 db.

Several methods of synchronizing the local oscillator with the incoming carrier are in use. For the most complex receivers, the received carrier is separated out and used to drive a servomechanism that mechanically adjusts the oscillator variable condenser. This allows extremely good control but is also very expensive. This class of receiver will not be discussed in this paper. Another method of synchronization control is the use of a reactance tube, which is controlled by the D.C. component of the carrier out of the demodulator. This reactance tube, in turn, by providing a variable capacitive reactance shunted across the local oscillator tank circuit, tends to synchronize the local oscillator.



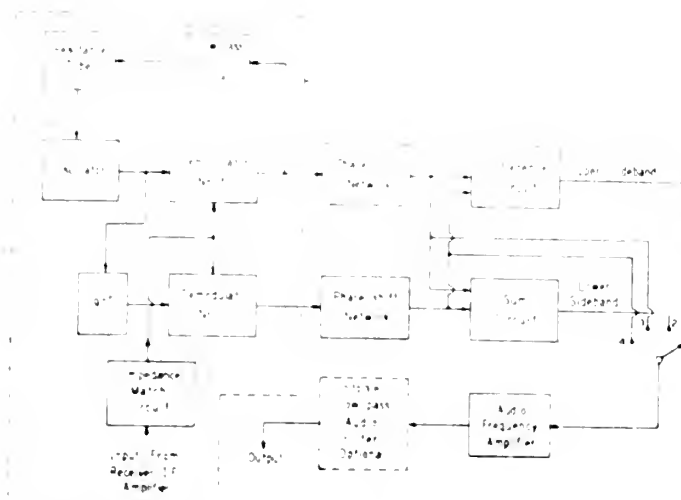
There is room for much development work in the field of relatively simple synchronizing devices.

One of the advantages of SSSC is that as half the frequency spectrum is required as compared with A.M., the ratio of signal to noise will be doubled in a receiver that is selective enough to accept only the usable sideband spectrum. This desirable feature may be realized from an A.M. transmission if the receiver is capable of removing one sideband. By using the phasing methods described under SSSC transmission, an adapter may be designed that will permit the removal of either sideband in a standard A.M. transmission. Into this converter can be incorporated the following features:

- (a. Single sideband reception of A.M. transmissions
- (b. Exalted carrier reception of A.M. transmissions
- (c. SSSC reception (either sideband)
- (d. Automatic frequency synchronization on either  
A.M. carrier, or SSRC transmissions.

Figure 12 is a block diagram of such an adapter. Consider first Figure 13, which is the enclosed portion of Figure 12, neglecting the low-pass R-C filter and the reactance tube. The input from the receiver I.F., which would have gone into the receiver detector, is fed into two identical demodulator tubes. A local oscillator, tuned to carrier frequency, feeds directly into one of the demodulator tubes, and through a  $90^\circ$





**FIGURE 12.** A Block Diagram of a Single-Sideband Receiving System.<sup>15</sup>









phase-shifter, into the other. The amplitude of this local oscillator signal is approximately 100 volts, thus it controls the operation of each demodulator to give exalted-carrier demodulation of A.M. If an incoming signal is of somewhat different frequency than the local oscillator, then a beat note will be produced. Each demodulator will put out this beat note, let us say of 1000 cps. The frequency and amplitude will be the same for the two demodulators, however, there will be a  $90^\circ$  phase shift between the two. Further, this  $90^\circ$  phase shift will be either positive or negative, depending on whether the incoming signal is of higher or lower frequency than that of the local oscillator. These two audio signals leave the demodulators and pass through two phase-shift networks that together, produce a further phase shift of  $90^\circ$  in a fixed direction. Let us assume that the incoming wave was 1000 cps. higher than the local oscillator frequency. This would result in two 1000 cps audio signals from the two demodulators that differ in phase by  $90^\circ$ . Now, these two audio signals are fed into the phase-shift networks that introduce an additional phase lag of  $90^\circ$ . The two audio signals thus emerge from the phase-shift networks  $180^\circ$  apart. If the two are added, zero audio signal results.

If the two are subtracted, then maximum signal will result.



Suppose now, that in addition to the incoming wave of 1000 cps greater frequency, a wave of 1000 cps less than the local oscillator is also fed into the demodulators. This is equivalent to an A.M. wave modulated at 1000 cps. Then the higher frequency will result in the same output as described above. The lower frequency, however, will result in audio signals out of the demodulators that lead each other by  $90^\circ$ . When these are passed through the phase shifters, the output signals will be exactly in phase. Thus, in a "sum" circuit, the 1000 cps note due to the lower incoming wave (lower sideband) alone would be present. In a "difference" circuit, the 1000 cps note due to the higher frequency wave (upper sideband) alone would be present. The two sidebands have been effectively split, and reception is possible on either alone. Of course, if only one sideband is being transmitted, then the audio signal will appear at either the "sum" or the "difference" circuit, but not at both.

The low-pass R-C filter permits only the D.C. component of the first demodulator to enter the reactance tube, which, as already mentioned, serves to synchronize the local oscillator with the incoming carrier.

In order to appreciate the advantage of single sideband reception of A.M. waves and the use of exalted carrier detection, some discussion should be given to the problems of



radio communications. This does not include commercial broadcasting, as in most cases, sufficient separation is provided either geographically or in transmission frequency to prevent interference. On short wave communications frequencies, however, geography is of little help, and the number of stations is very much greater. The frequency separation between stations may approach zero. The amateur channels are excellent examples of the situation being discussed. Very narrow bands are provided, and thousands of amateurs attempt to say their bit at the same time. The condition may be expected to be as bad on commercial communications channels in a few years. If a receiver is tuned to one station and another, removed in frequency by 5 KC (which is very common) is transmitting, then a 5 KC audio beat note will result in a conventional receiver. In actual practice, this is the most common type of interference present, not only on the amateur bands, but also on many of the communications channels. (If suppressed carrier transmission were employed, this disturbance would be reduced to a negligible value). The use of an adapter herein described would permit the receiver operator to choose the particular sideband that was least disturbed by these interfering heterodynes. A further difficulty encountered is the blanketing effect of a strong transmitter on a relatively weak station if





the two are not far removed in frequency. The use of an exalted local carrier, accurately tuned to the weaker station frequency, would cause the detector to reject the strong station in favor of the desired weaker station. A further advantage of exalted carrier detection is also present. In detection of an A.M. wave as received, the value of  $M$  must not exceed unity or very great distortion will result. This is the same type of distortion that would be present if a transmitter were modulating greater than 100%. Even if the transmitter has an  $M$  of less than unity, frequently selective fading will reduce the relative magnitude of the carrier. This may be noticed at times when receiving a broadcast station that is some distance away, particularly at dusk, when a shift of the reflected ionized layer is occurring. The use of a strong locally inserted carrier greatly reduces difficulties of this type.

In the adapter described here, there are no inherent limitations on the width of the audio band as is the case for other highly selective circuits. However, for communications use, an audio frequency range of from approximately 200 cps to 2800 cps is adequate for satisfactory intelligence. If an audio filter is employed at the transmitter, appreciable savings in power result from the blocking of frequencies outside this range. At the receiving end, the effective selectivity of the



receiver may be greatly improved, and effectively a nearly ideal rectangular selectivity curve is obtained, as is shown in Figure 14. In Appendix VI are given the details of construction of an adapter utilizing the principles outlined above. In order to gain first-hand experience with construction and application of such a unit, this adapter was constructed. Operation is very satisfactory. A reduction in the unwanted sideband of 35 db was obtained.



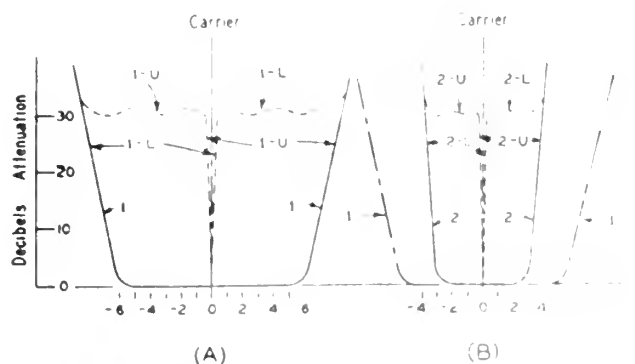


Fig. 14 Over the range for which the audio phase-shift networks hold close to  $90^\circ$  difference, the apparent i.f. response of the system is determined in combination by the i.f. and audio bandwidths. The characteristic with no audio filter is shown at (A), and (B) demonstrates how greater effective selectivity is obtained with an audio filter.



CHAPTER IV  
ADVANTAGES, DISADVANTAGES AND LIMITATIONS OF  
SSSC AND SSRC

Considerable confusion can arise in the consideration of advantages of SSSC transmission because of the variety of viewpoints possible. That very definite advantages do exist is not subject to dispute, but the degree of the advantages depends on what is desired of the transmitter. Figure 15 represents a conventional A.M. transmitter. Say for example,

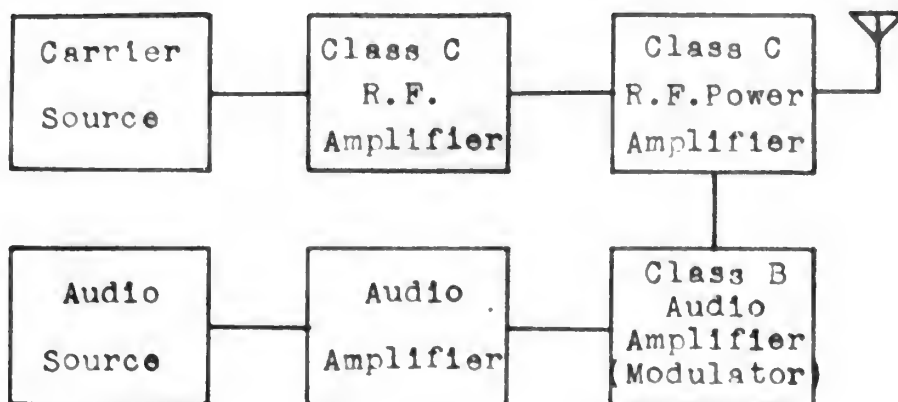


Figure 15. Block Diagram of a Conventional A.M. Transmission System.

it is rated at 10,000 watts. This is the power output of the R.F. power amplifier with no modulation. As the efficiency of a good Class C.R.F. amplifier is about 80%, the power input to the stage will be approximately the 12,500 watts. The required audio power for 100% modulation will be half





of the power input to the final R.F. amplifier, or 6250 watts. At 100% modulation, the average power radiated by the transmitter is  $.80(12,500 + 6250)$ , or 15,000 watts, of which only 5000 watts is useful in the transmission of intelligence. Now, an often overlooked point is that for a normal human voice, the ratio of average to peak power is never larger than  $1/2$ , and it is usually  $1/4$  or less. Then for a peak audio power of 5000 watts, the average is only 1250 watts. Thus for a 10,000 transmitter, the average power radiated is 11,250 watts, and the peak power is (four times the carrier power) 40,000 watts. Amplifier tubes are rated in size by their power dissipating ability. In this hypothetical transmitter, at no modulation, average dissipated power is 2500 watts. At 100% modulation on peaks of normal voice, the average dissipated power is 2810 watts, and the peaks of dissipated power is 10,000 watts. Thus the tube in the final R.F. amplifier would have to have a power rating of 2810 watts and would have to be able to withstand peaks of 10,000 watts.

From this point, two different paths may be followed. First, consider that it is only desired to transmit an average of 1250 watts, with a peak of 5000 watts of intelligence -- the same as for the original transmitter -- by SSSC transmission.



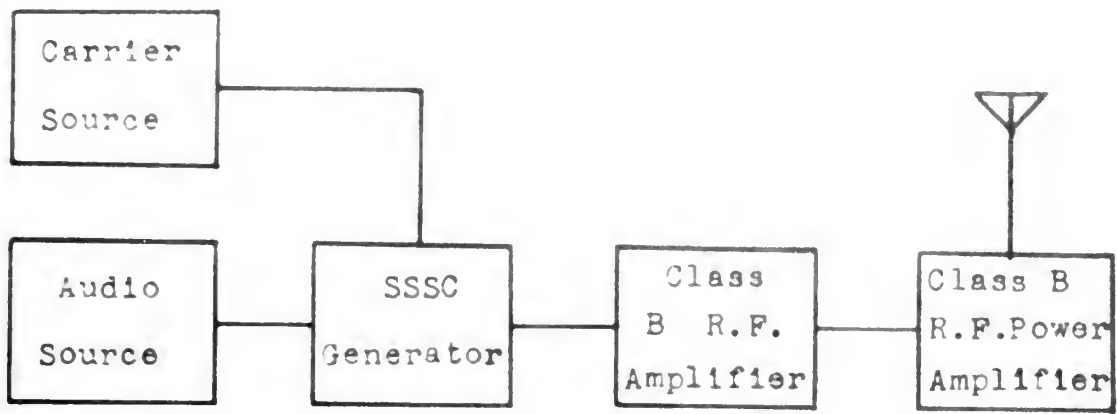


Figure 16. Block Diagram of SSSC Transmission System

For SSSC, this power represents the entire output. There is no carrier. Further, this peak of 5000 watts of intelligible power is the peak power of the transmitter, because for uniform intensity of audio signal, the envelope of a SSSC wave is a pair of horizontal lines. Note, however, that linear R.F. amplifiers must be used for SSSC, and the efficiency of a good Class B stage is about 70%. Thus, the average power input to the final R.F. amplifier is  $1250/.70$ , or 1786 watts; and the dissipated power is 535 watts. The peak of power dissipation in the final amplifier is 2140 watts.



In Figure 16, it is important to note that all stages up to the first R.F. amplifier are of negligible power. Receiver-type elements are employed. The only power stages are the R.F. amplifiers. No audio power amplifiers are required. Comparing Figures 15 and 16 in this light, it may be estimated that roughly 25 KW of line power is required for the A.M. transmitter, while approximately 8 KW of line power is required by the SSSC transmitter.

For a SSSC transmitter designed to transmit an equal amount of intelligible power, the advantages may be summarized as follows:

- (a. Savings of approximately 80% in size and cost of final amplifier.
- (b. Savings in cost, and reduction in size of transmitter, by complete elimination of audio power amplifiers and power modulators.
- (c. Savings of approximately 70% in cost of operating power.

No significant disadvantages or limitations have yet been introduced.

Suppose, now, that the original A.M. transmitter is on hand, and it is desired to realize all possible output power advantages by conversion to SSSC. What, then,



would be the benefits to be derived. Reviewing, the A.M. transmitter contains a Class C final amplifier, capable of dissipating an average of 2812.5 watts, with a dissipation peak of 10,000 watts. Operated as a Class B, SSSC final amplifier, this would be able to deliver 9370 watts, average, of intelligible power, with peaks of 33,333 watts. As this ratio is greater than the estimated 1 to 4, using the peak power as the limiting criterion, an average power of 8320 watts could be transmitted. As before, the audio power amplifiers are eliminated, and operating power consumption is reduced to approximately 13 KW.

Summarizing the advantages to be obtained by converting an existing A.M. transmitter to SSSC to obtain greatest possible output power, they are:

- (a. Intelligible power transmitted increased by approximately 650%.
- (b. Audio power amplifiers eliminated.
- (c. Savings of approximately 50% in cost of operating power.

Again, other than the work of converting an existing transmitter, no appreciable disadvantages or limitations exist.





Before leaving the transmission end of SSSC, mention should be made of several practical items that require attention. First, the power supply to all stages of a SSSC transmitter must be extremely well filtered to prevent spurious radiations and prevent undue distortion of the audio signal. Next, all stages must be carefully designed and operated, so as to be as nearly linear as possible, otherwise, the magnitude of the suppressed sideband will become appreciable due to the generation of an A.M. signal in the non-linear elements. In general, more care and attention to details of design is required in a SSSC transmitter if really satisfactory operation is to be obtained. These items are problems of design, construction and operation and should not be considered as disadvantages or limitations of the system.

Advantages and limitations inherent in single sideband reception of either A.M. or SSSC transmissions have been adequately discussed. The limitations and most of the disadvantages of the SSSC system lie in the reception end. Summarizing these from the discussion of SSSC and SSRC reception, they are:

Disadvantages--

- (a. Reception impossible on all but communications-type receiver.



- (b. Reception procedure more complex, and in some cases, impossible on conventional communications receivers. Results are only fair.
- (c. Even with moderately priced adaptors, the problem of tuning is more complex than for A.M. Simple automatic lock-in circuits are not thoroughly developed (assuming a reduced carrier is transmitted).
- (d. Everpresent problem of satisfactory local oscillator stability is difficult to satisfy without either loss of flexibility of operation or high cost.

Advantages —

- (a. Increase in signal-to-noise ratio.
- (b. Ability to choose sideband having least interfering signals.
- (c. Ability to better separate desired weak signal from undesirable strong signal.

Some of the advantages of the SSSC system can not be assigned to either the transmitter or the receiver alone. Of greatest importance in this classification is the fact that only half the R.F. spectrum is required for SSSC as compared with A.M. This permits either a wider separation of stations, or a larger number of simultaneously operating



stations. <sup>The</sup> second, is a result on reception of suppressing carrier. As mentioned before, on communications channels, one of the greatest sources of interference is the productions of very intense hetrodyne notes as a result of mixing of carriers not far removed in frequency. Elimination of carriers would, of course, completely eliminate this disturbance. Further, as no carrier is transmitted, if the transmitter is operating properly, when the operator is not speaking, no signal at all is transmitted (except possibly a greatly reduced carrier, which would not effect the results.) This immediately permits breakin operation with no special equipment, and on the same frequency.

In view of the above discussion, several conclusions seem reasonable. First, use of SSSC on commercial broadcast bands, while desirable from the point of view of station economy and efficiency of frequency utilization, is not feasible because of the handicaps of reception. Not only would converters be required for the 80 million home receivers now in use, but also, an educational program in tuning procedure would be required. This is ridiculous, of course.

For commercial communications operation, however, the story is entirely different. Except for a few classes of communications, the use of A.M. is not only extremely



wasteful of equipment and power, but the waste of 50% of the available frequency is hardly excusable with SSSC in it's present state of development. The number of communications stations is increasing rapidly with time. For a long while, the demand for additional frequency space was met (largely due to the pioneering of the amateur radio operators) by going to higher and higher frequencies. With the spectrum of communications circuits approaching the frequencies of light, the limit seems to be in sight. Further, the communications circuits for which SSSC is most applicable are not satisfactory at frequencies much above 30 megacycles. The available space, then, seems to be from 50 to 550 KC and from 1.7 to 30 MC.

The expense of conversion of a transmitter would not be prohibitive. Even this cost could be more than regained by savings in operating power costs over a reasonable period of time. To the writer, it seems only reasonable to expect that more and more newly constructed communications stations will be of the SSSC variety, and that conversion of existing transmitters will accelerate as operating procedure knowledge is more widespread. It does not seem unreasonable to expect that in the not too distant future, the great majority of communications links will employ SSSC systems. For amateur operation, the present situation





in the "phone" bands is a good example of what is to be expected in commercial channels as the number of stations increases -- unless SSSC is more universally adopted. Perhaps if SSSC comes into widespread use in the amateur channels, not only will the bad situation be cleared up within those channels, but as has happened in the past, developments of technique and simplifications of equipment will cause a more rapid spread of SSSC into the communications channels.



## BIBLIOGRAPHY

1. Caruthers, R. S. Copper Oxide Modulators in Carrier Telephone Systems.  
Bell System Technical Journal, April 1939.
2. Terman, F. E. Radio Engineering.  
McGraw-Hill Book Company, Inc., 1947.
3. Reich, H. J. Theory and Applications of Electron Tubes.  
McGraw-Hill Book Company, Inc., 1939.
4. Dome, R. B. Wideband Phase Shift Networks.  
Electronics, December 1946.
5. Honnell, M. A. Single-Sideband Generator.  
Electronics, November 1945.
6. Crosby, M. G. Exalted Carrier Amplitude and Phase Modulation Reception.  
Proceedings, IRE Vol. 32, No. 9, September 1945.
7. McLaughlin, J. L. A. Exit Hetrodyne QRM.  
QST, October 1947.
8. Goodman, Byron What is Single-Sideband Telephony?  
QST, January 1948.
9. Villard, Jr., O. G. Single-Sideband Operating Tests.  
QST, January 1948.
10. Nichols, A. H. A Single-Sideband Transmitter for Amateur Operation.
11. McLaughlin, J. L. A. Selectable Single-Sideband Reception Simplified.  
QST, April 1948.
12. Villard, Jr., O. G. Selectivity in SSSC Reception.  
QST. April 1948.
13. Norgaard, D. E. What About Single-Sideband?  
QST, May 1948.



14. Norgaard, D. E. A New Approach to Single-Sideband.  
QST, June 1948.
15. Norgaard, D. E. Practical Single-Sideband Reception.  
QST, July 1948.
16. Villard, O. G. & Thompson, D. L. A Detector for  
Single-Sideband Reception.  
QST, June 1948.
17. G. E. Ham News, Nov.-Dec. 1948.
18. Heising, R. A. Production of Single-Sideband for  
Transatlantic Radio Telephony.  
Proc. IRE; Vol. 13, No. 6, June 1925.
19. Reque, S. G. Linear R. F. Amplifiers.  
QST, May 1949.
20. E. E. Staff, M.I.T., Applied Electronics.  
John Wiley & Sons, Inc., 1946.
21. Everitt, W. L. Communication Engineering.  
McGraw-Hill Book Company, 1937.
22. Berry, F. M. A Filter Design for the Single-Sideband  
Transmitter.  
QST, June 1949.
23. Herold, E. W., and Malter, L. Some Aspects  
of Radio Reception at Ultra-High Frequencies,  
Part V. Frequency mixing in Diodes.  
Proc. IRE; Vol. 31, No. 10, October 1943.
24. Herold, E. W. The Operation of Frequency Converters  
and Mixers for Superhetrodyne Reception.  
Proc. IRE; Vol. 30, No. 2, February 1942.
25. Arrington, II, J. L. Germanium Varistors (Thesis).



# A Filter Design for the Single-Sideband Transmitter

BY IRLE M. BERRY \* W6MNN

**S**INGLE-SIDE-BAND (SSB) transmitting systems have been described in the literature for many years. The first practical SSB transmitter was described by V. F. Zentgraf in 1923. Since that time, many different methods have been proposed for the generation of SSB signals. The most common method is the use of a balanced modulator. This method is simple and efficient, but it requires a high degree of balance in the modulator. The second method is the use of a phase shifter. This method is also simple, but it requires a high degree of phase balance. The third method is the use of a filter. This method is the most complex, but it is the only method that can be used to generate SSB signals with a high degree of balance and phase balance.

Briefly, the method of generating SSB signals by the use of a filter is as follows: A carrier wave is generated and a speech signal is applied to a balanced modulator. The output of the modulator is a double-sideband signal. This signal is then passed through a filter which removes one of the sidebands. The remaining sideband is the SSB signal. The filter must be designed to pass the desired sideband and reject the other sideband. This is a difficult task, but it can be accomplished by the use of a high-pass or low-pass filter.

\* 1200 E. 4th Terrace, Kansas City 4, Mo.  
 † Goodman, "What Is Single-Sideband Telephony?" QST, Jan. 1948. Norgess, "What About Single-Sideband?" QST, May 1948.  
 ‡ Norgess, "A New Approach to Single-Sideband," QST, June 1948.

The filter method of generating SSB signals is the most complex, but it is the only method that can be used to generate SSB signals with a high degree of balance and phase balance. The filter must be designed to pass the desired sideband and reject the other sideband. This is a difficult task, but it can be accomplished by the use of a high-pass or low-pass filter. The filter must be designed to pass the desired sideband and reject the other sideband. This is a difficult task, but it can be accomplished by the use of a high-pass or low-pass filter.

The filter method of generating SSB signals is the most complex, but it is the only method that can be used to generate SSB signals with a high degree of balance and phase balance. The filter must be designed to pass the desired sideband and reject the other sideband. This is a difficult task, but it can be accomplished by the use of a high-pass or low-pass filter. The filter must be designed to pass the desired sideband and reject the other sideband. This is a difficult task, but it can be accomplished by the use of a high-pass or low-pass filter.

The block diagram of a practical single-sideband transmitter is shown in Fig. 1. The selection of upper or lower sideband is accomplished by switching the frequency of the second oscillator. In the notation of Fig. 1, "USB" and "LSB" indicate the position of sideband at the points noted with respect to the input speech frequencies. This is not to be confused with the particular sideband of the second oscillator that is selected by the second IF. Careful study of Fig. 1 will make this clear.

Although not directly indicated in Fig. 1, the

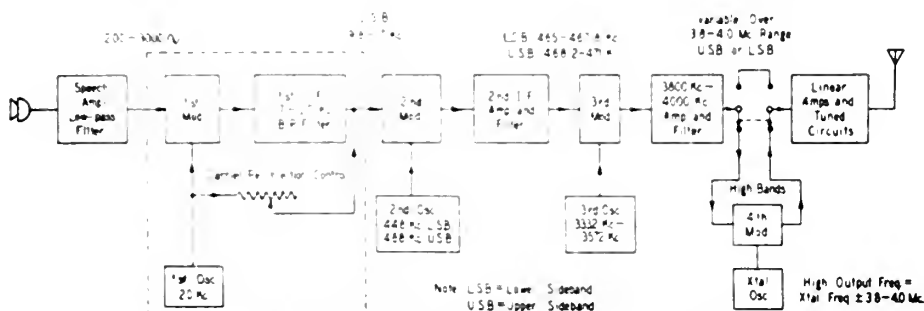


Fig. 1—Block diagram of a typical single-sideband, suppressed-carrier transmitter or exciter. The equipment inside the dashed area is described in this article.

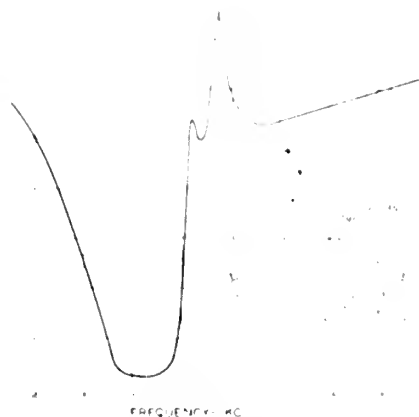




Except in the case of  $C_1$  and  $C_2$  (Fig. 1), mica or other low-loss types of capacitors are necessary for proper filter action.  $C_1$ ,  $C_2$  and  $C_3$  are large values for mica capacitors and would be expensive. Ordinarily they would have to be made from a large number of paralleled units.  $C_1$  and  $C_2$  appear directly in parallel with the terminating resistances and it has been found that good-quality paper capacitors are satisfactory here.



Fig. 4 — Component values of the filter after combination of the parallel components. Values are in  $\mu$ d. and mh.



$C_1$  is located internally and must be of low-loss type for best results. It is possible, however, to use impedance transformation at  $L_1/C_1$  and permit a smaller mica capacitor to be used for  $C_1$  at the expense of a large  $Q$  value for  $L_1$ . The method of impedance transformation employed also permits a relaxation of the capacitor tolerance. Any reasonable value may be used for  $C_1$  provided the inductor is adjusted to the correct anti-resonant frequency. The correct impedance may then be regained by tapping  $L_2$ . With inductors of high coupling between turns, the proper point of tapping is such that the inductance between common and tap is approximately that of the value of  $L_2$  before transformation. The modification with the more desirable values is shown in Fig. 6.

In the filter of Fig. 3 the tolerance of some of the elements is quite critical, particularly that of the series arms. It has been found in the design of filters of this type that the tolerance of  $LC$  ratios is not particularly critical provided the correct resonant and antiresonant frequencies are maintained. Practically, this leads to the selection of capacitors to a tolerance of  $\pm 5$  per cent or better, and resonating each  $LC$  circuit to the correct frequency by turn adjustment of the inductor. The maximum possible error of 10 per cent in the impedance match between junctions of the filter arms is not serious. Greater tolerances will cause a "ripple" in the passband and other deviations from the desired characteristics. In following this procedure, note that the series arms of the filter have both a resonant and an antiresonant frequency, with the inductor as a common element for both. Obviously, the inductor could not be adjusted







independent of the value of  $C_2$ , thus desired independent adjustment of inductor arrangement is used.

Considering the series combination of  $L_2$  and  $C_2$  selected with a definite value of  $C_2$ , the inductance  $L_2$  may then be adjusted with this series combination to the correct antiresonant frequency. It may then have fewer turns than the original calculated value of  $L_2$ . Leaving  $C_2$  connected, to pass the exact number of turns necessary for antiresonance, turns may be added to  $L_2$  until the series combination of  $L_2$ ,  $C_2$  in series with  $C_1$  and the added winding of  $L_2$  will series resonate at the correct frequency. The exact value of  $C_2$  will set the impedance of the circuit and a tolerance of 5 percent is permissible.

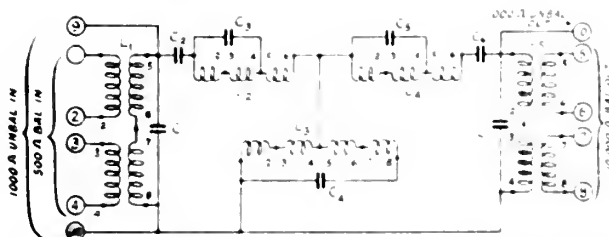


Fig. 6—Revised filter of Fig. 4, with provision for balanced or unbalanced input and output.

- $C_1$ —0.049  $\mu$ f.  $\pm 5\%$ , paper
- $C_2$ —0.0015  $\mu$ f.  $\pm 5\%$ , mica
- $C_3$ —0.03  $\mu$ f.  $\pm 10\%$ , mica
- $C_4$ —0.03  $\mu$ f.  $\pm 20\%$ , mica
- $C_5$ —0.011  $\mu$ f.  $\pm 10\%$ , mica
- $C_6$ —0.0031  $\mu$ f.  $\pm 5\%$ , mica
- $C_7$ —0.051  $\mu$ f.  $\pm 5\%$ , paper
- $L_1$  (1-4)—0.7 mh., 33 turns No. 26, bifilar
- $L_2$  (5-8)—1.4 mh., approx. 47 turns No. 26, bifilar
- $L_3$  (1-4)—2.0 mh., approx. 80 turns No. 26, bifilar
- $L_4$  (5-6)—0 to 20 turns No. 26, single
- $L_5$  (1-4)—0.7 mh., approx. 47 turns No. 26, bifilar
- $L_6$  (1-8)—2.3 mh., approx. 86 turns No. 26, bifilar
- $L_7$  (1-4)—5.0 mh., approx. 125 turns No. 26, bifilar
- $L_8$  (5-6)—0 to 20 turns No. 26, single
- $L_9$  (5-8)—14.0 mh., 160 turns No. 28, bifilar
- $L_{10}$  (1-4)—1.4 mh., approx. 47 turns No. 26, bifilar

All wire Formvar or d.s.c.—see text.  
 $L_1$  and  $L_5$  wound on Western Electric P170930 core ring.  
 $L_2$ ,  $L_3$  and  $L_4$  on Western Electric P281195 core ring.

$$\text{Approx. turns P470930} = 1000 \sqrt{\frac{164}{f}}$$

$$\text{Approx. turns P281195} = 1000 \sqrt{\frac{79}{f}}$$

Resonant Frequencies		
Inductor	Capacitor	Freq., k.
$L_1$ (5-8)	$C_1$	19.1
$L_2$ (1-4)	$C_2$	20.5
$L_3$ (1-6) with $C_2$ connected	$C_2$	19.1
$L_4$ (1-4)	0.1 $\mu$ f.	19.0
	$\pm 5\%$	
$L_2$ (1-8)	$C_4$	19.0
$L_4$ (1-4)	$C_5$	21.5
$L_6$ (1-6) with $C_5$ connected	$C_5$	18.9
$L_8$ (1-4)	$C_7$	18.9

$$L = \frac{1}{4\pi^2 f^2 C} \quad (1)$$

For a series combination of  $L_2$  and  $C_2$  in parallel operation, the value of  $L_2$  and  $C_2$  will be the same as the value of  $L_2$  and  $C_2$  in series operation. For a parallel combination of  $L_2$  and  $C_2$  in series operation, the value of  $L_2$  and  $C_2$  will be the same as the value of  $L_2$  and  $C_2$  in parallel operation.

In the design of a filter of impedance of 5000 ohms, the value of  $C_2$  was selected at 0.0015  $\mu$ f. This value is a compromise between desired voltage step-up and keeping the number of turns to a value that permits hand winding. The completed filter design after all modifications is shown in Fig. 6. In the event other input or output impedances are desired, the number of turns and method of connection of these added windings may be altered to meet the requirements. Since the impedance varies directly with the inductance of the windings, with 1.4 mh. the inductance for 1000 ohms impedance, the required inductance in millihenrys for any new impedances may be found by dividing the new impedance in ohms by 1000 and multiplying the result by 1.4. The number of turns required can be found from the formulas for the inductors given in Fig. 6.

### Filter Alignment

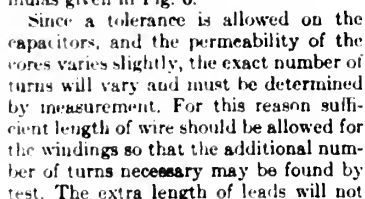
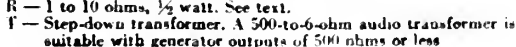
As has previously been mentioned, the LC combinations must be resonant at the desired frequency. In an m-type filter with closely-spaced rejection frequencies, it is very important to hold to very close frequency tolerances, while a constant error is not serious the spacing of one frequency to the next is critical.

Heretofore, it has been considered necessary to use expensive laboratory equipment, which is out of the reach of many. Signal generators for the range of 15 to 30 kc. are not common, and those available are usually not of sufficient accuracy. However, with the aid of a BC-221 frequency meter the main obstacle has been removed. The fundamental frequency range of the low band of the BC-221 is 125-250 kc., and it has sufficient output voltage to give a reasonable indication on most oscilloscopes. The BC-221 is used only as a calibration means for the test signal generator. The test generator may easily be made from the junk box, and the usual calibration problem



A question arises: is assuming that  $\beta$  is Gaussian (as in the results shown in Fig. 7) the most reasonable? The answer is not clear. It is not clear whether the  $1/f$  correlations in  $\beta$  are as strong as indicated in Fig. 7. With the  $1/f$  correlation model we proceed as shown in Fig. 8, a sharp increase in impedance occurs when the frequency is at the exact sensor resonance point. Since the impedance is then a minimum (Fig. 8), although the signal  $R$  of Fig. 7 should be the  $1/f$  values, a dip that will still give a noticeable indication. When an entire sensor arm is used, the dip will not be as great but will be a sharp dip.

In selecting indicators for the filter, the  $Q$  is of primary importance.  $Q$  values of at least 150 are necessary for all indicators except possibly  $I_1$  and  $I_2$ . Ignored  $I_3$ , as in the case of  $I_4$  and  $I_5$ , are in parallel with the formants, and less so here, have much less effect. While many types of indicators might be used, the toroidal type has many advantages and core rings of molybdenum. Personnel are now available to the supplier.







affect the test, and these lead wires are cut to proper length after the correct number of turns has been determined.

Wire size is not critical and depends on the that given in Fig. 6 may be made it with wire that does not limit. "Formvar" insulation or the equivalent, is recommended and is easy to wind, but single silk or nylon is satisfactory. Plain enamel, not of the Formvar type is to be avoided, because of the possibility of shorted turns.

In order to reduce the number of times the wire must be threaded through the core ring, all windings are bifilar except the adjustment windings 5-6 of  $L_2$  and  $L_4$ . In the bifilar type of winding, two wires are held together and wound as one. After winding, the start of one wire (3) may be connected to the finish of the other (2), thus connecting the two in the series-aiding manner. As in telephone practice, the numbering of

windings is as follows: 1-2, 3-4, 5-6, 7-8, 9-10, 11-12, 13-14, 15-16, 17-18, 19-20, 21-22, 23-24, 25-26, 27-28, 29-30, 31-32, 33-34, 35-36, 37-38, 39-40, 41-42, 43-44, 45-46, 47-48, 49-50, 51-52, 53-54, 55-56, 57-58, 59-60, 61-62, 63-64, 65-66, 67-68, 69-70, 71-72, 73-74, 75-76, 77-78, 79-80, 81-82, 83-84, 85-86, 87-88, 89-90, 91-92, 93-94, 95-96, 97-98, 99-100, 101-102, 103-104, 105-106, 107-108, 109-110, 111-112, 113-114, 115-116, 117-118, 119-120, 121-122, 123-124, 125-126, 127-128, 129-130, 131-132, 133-134, 135-136, 137-138, 139-140, 141-142, 143-144, 145-146, 147-148, 149-150, 151-152, 153-154, 155-156, 157-158, 159-160, 161-162, 163-164, 165-166, 167-168, 169-170, 171-172, 173-174, 175-176, 177-178, 179-180, 181-182, 183-184, 185-186, 187-188, 189-190, 191-192, 193-194, 195-196, 197-198, 199-200, 201-202, 203-204, 205-206, 207-208, 209-210, 211-212, 213-214, 215-216, 217-218, 219-220, 221-222, 223-224, 225-226, 227-228, 229-230, 231-232, 233-234, 235-236, 237-238, 239-240, 241-242, 243-244, 245-246, 247-248, 249-250, 251-252, 253-254, 255-256, 257-258, 259-260, 261-262, 263-264, 265-266, 267-268, 269-270, 271-272, 273-274, 275-276, 277-278, 279-280, 281-282, 283-284, 285-286, 287-288, 289-290, 291-292, 293-294, 295-296, 297-298, 299-300, 301-302, 303-304, 305-306, 307-308, 309-310, 311-312, 313-314, 315-316, 317-318, 319-320, 321-322, 323-324, 325-326, 327-328, 329-330, 331-332, 333-334, 335-336, 337-338, 339-340, 341-342, 343-344, 345-346, 347-348, 349-350, 351-352, 353-354, 355-356, 357-358, 359-360, 361-362, 363-364, 365-366, 367-368, 369-370, 371-372, 373-374, 375-376, 377-378, 379-380, 381-382, 383-384, 385-386, 387-388, 389-390, 391-392, 393-394, 395-396, 397-398, 399-400, 401-402, 403-404, 405-406, 407-408, 409-410, 411-412, 413-414, 415-416, 417-418, 419-420, 421-422, 423-424, 425-426, 427-428, 429-430, 431-432, 433-434, 435-436, 437-438, 439-440, 441-442, 443-444, 445-446, 447-448, 449-450, 451-452, 453-454, 455-456, 457-458, 459-460, 461-462, 463-464, 465-466, 467-468, 469-470, 471-472, 473-474, 475-476, 477-478, 479-480, 481-482, 483-484, 485-486, 487-488, 489-490, 491-492, 493-494, 495-496, 497-498, 499-500, 501-502, 503-504, 505-506, 507-508, 509-510, 511-512, 513-514, 515-516, 517-518, 519-520, 521-522, 523-524, 525-526, 527-528, 529-530, 531-532, 533-534, 535-536, 537-538, 539-540, 541-542, 543-544, 545-546, 547-548, 549-550, 551-552, 553-554, 555-556, 557-558, 559-560, 561-562, 563-564, 565-566, 567-568, 569-570, 571-572, 573-574, 575-576, 577-578, 579-580, 581-582, 583-584, 585-586, 587-588, 589-590, 591-592, 593-594, 595-596, 597-598, 599-600, 601-602, 603-604, 605-606, 607-608, 609-610, 611-612, 613-614, 615-616, 617-618, 619-620, 621-622, 623-624, 625-626, 627-628, 629-630, 631-632, 633-634, 635-636, 637-638, 639-640, 641-642, 643-644, 645-646, 647-648, 649-650, 651-652, 653-654, 655-656, 657-658, 659-660, 661-662, 663-664, 665-666, 667-668, 669-670, 671-672, 673-674, 675-676, 677-678, 679-680, 681-682, 683-684, 685-686, 687-688, 689-690, 691-692, 693-694, 695-696, 697-698, 699-700, 701-702, 703-704, 705-706, 707-708, 709-710, 711-712, 713-714, 715-716, 717-718, 719-720, 721-722, 723-724, 725-726, 727-728, 729-730, 731-732, 733-734, 735-736, 737-738, 739-740, 741-742, 743-744, 745-746, 747-748, 749-750, 751-752, 753-754, 755-756, 757-758, 759-760, 761-762, 763-764, 765-766, 767-768, 769-770, 771-772, 773-774, 775-776, 777-778, 779-780, 781-782, 783-784, 785-786, 787-788, 789-790, 791-792, 793-794, 795-796, 797-798, 799-800, 801-802, 803-804, 805-806, 807-808, 809-810, 811-812, 813-814, 815-816, 817-818, 819-820, 821-822, 823-824, 825-826, 827-828, 829-830, 831-832, 833-834, 835-836, 837-838, 839-840, 841-842, 843-844, 845-846, 847-848, 849-850, 851-852, 853-854, 855-856, 857-858, 859-860, 861-862, 863-864, 865-866, 867-868, 869-870, 871-872, 873-874, 875-876, 877-878, 879-880, 881-882, 883-884, 885-886, 887-888, 889-890, 891-892, 893-894, 895-896, 897-898, 899-900, 901-902, 903-904, 905-906, 907-908, 909-910, 911-912, 913-914, 915-916, 917-918, 919-920, 921-922, 923-924, 925-926, 927-928, 929-930, 931-932, 933-934, 935-936, 937-938, 939-940, 941-942, 943-944, 945-946, 947-948, 949-950, 951-952, 953-954, 955-956, 957-958, 959-960, 961-962, 963-964, 965-966, 967-968, 969-970, 971-972, 973-974, 975-976, 977-978, 979-980, 981-982, 983-984, 985-986, 987-988, 989-990, 991-992, 993-994, 995-996, 997-998, 999-1000, 1001-1002, 1003-1004, 1005-1006, 1007-1008, 1009-1010, 1011-1012, 1013-1014, 1015-1016, 1017-1018, 1019-1020, 1021-1022, 1023-1024, 1025-1026, 1027-1028, 1029-1030, 1031-1032, 1033-1034, 1035-1036, 1037-1038, 1039-1040, 1041-1042, 1043-1044, 1045-1046, 1047-1048, 1049-1050, 1051-1052, 1053-1054, 1055-1056, 1057-1058, 1059-1060, 1061-1062, 1063-1064, 1065-1066, 1067-1068, 1069-1070, 1071-1072, 1073-1074, 1075-1076, 1077-1078, 1079-1080, 1081-1082, 1083-1084, 1085-1086, 1087-1088, 1089-1090, 1091-1092, 1093-1094, 1095-1096, 1097-1098, 1099-1100, 1101-1102, 1103-1104, 1105-1106, 1107-1108, 1109-1110, 1111-1112, 1113-1114, 1115-1116, 1117-1118, 1119-1120, 1121-1122, 1123-1124, 1125-1126, 1127-1128, 1129-1130, 1131-1132, 1133-1134, 1135-1136, 1137-1138, 1139-1140, 1141-1142, 1143-1144, 1145-1146, 1147-1148, 1149-1150, 1151-1152, 1153-1154, 1155-1156, 1157-1158, 1159-1160, 1161-1162, 1163-1164, 1165-1166, 1167-1168, 1169-1170, 1171-1172, 1173-1174, 1175-1176, 1177-1178, 1179-1180, 1181-1182, 1183-1184, 1185-1186, 1187-1188, 1189-1190, 1191-1192, 1193-1194, 1195-1196, 1197-1198, 1199-1200, 1201-1202, 1203-1204, 1205-1206, 1207-1208, 1209-1210, 1211-1212, 1213-1214, 1215-1216, 1217-1218, 1219-1220, 1221-1222, 1223-1224, 1225-1226, 1227-1228, 1229-1230, 1231-1232, 1233-1234, 1235-1236, 1237-1238, 1239-1240, 1241-1242, 1243-1244, 1245-1246, 1247-1248, 1249-1250, 1251-1252, 1253-1254, 1255-1256, 1257-1258, 1259-1260, 1261-1262, 1263-1264, 1265-1266, 1267-1268, 1269-1270, 1271-1272, 1273-1274, 1275-1276, 1277-1278, 1279-1280, 1281-1282, 1283-1284, 1285-1286, 1287-1288, 1289-1290, 1291-1292, 1293-1294, 1295-1296, 1297-1298, 1299-1300, 1301-1302, 1303-1304, 1305-1306, 1307-1308, 1309-1310, 1311-1312, 1313-1314, 1315-1316, 1317-1318, 1319-1320, 1321-1322, 1323-1324, 1325-1326, 1327-1328, 1329-1330, 1331-1332, 1333-1334, 1335-1336, 1337-1338, 1339-1340, 1341-1342, 1343-1344, 1345-1346, 1347-1348, 1349-1350, 1351-1352, 1353-1354, 1355-1356, 1357-1358, 1359-1360, 1361-1362, 1363-1364, 1365-1366, 1367-1368, 1369-1370, 1371-1372, 1373-1374, 1375-1376, 1377-1378, 1379-1380, 1381-1382, 1383-1384, 1385-1386, 1387-1388, 1389-1390, 1391-1392, 1393-1394, 1395-1396, 1397-1398, 1399-1400, 1401-1402, 1403-1404, 1405-1406, 1407-1408, 1409-1410, 1411-1412, 1413-1414, 1415-1416, 1417-1418, 1419-1420, 1421-1422, 1423-1424, 1425-1426, 1427-1428, 1429-1430, 1431-1432, 1433-1434, 1435-1436, 1437-1438, 1439-1440, 1441-1442, 1443-1444, 1445-1446, 1447-1448, 1449-1450, 1451-1452, 1453-1454, 1455-1456, 1457-1458, 1459-1460, 1461-1462, 1463-1464, 1465-1466, 1467-1468, 1469-1470, 1471-1472, 1473-1474, 1475-1476, 1477-1478, 1479-1480, 1481-1482, 1483-1484, 1485-1486, 1487-1488, 1489-1490, 1491-1492, 1493-1494, 1495-1496, 1497-1498, 1499-1500, 1501-1502, 1503-1504, 1505-1506, 1507-1508, 1509-1510, 1511-1512, 1513-1514, 1515-1516, 1517-1518, 1519-1520, 1521-1522, 1523-1524, 1525-1526, 1527-1528, 1529-1530, 1531-1532, 1533-1534, 1535-1536, 1537-1538, 1539-1540, 1541-1542, 1543-1544, 1545-1546, 1547-1548, 1549-1550, 1551-1552, 1553-1554, 1555-1556, 1557-1558, 1559-1560, 1561-1562, 1563-1564, 1565-1566, 1567-1568, 1569-1570, 1571-1572, 1573-1574, 1575-1576, 1577-1578, 1579-1580, 1581-1582, 1583-1584, 1585-1586, 1587-1588, 1589-1590, 1591-1592, 1593-1594, 1595-1596, 1597-1598, 1599-1600, 1601-1602, 1603-1604, 1605-1606, 1607-1608, 1609-1610, 1611-1612, 1613-1614, 1615-1616, 1617-1618, 1619-1620, 1621-1622, 1623-1624, 1625-1626, 1627-1628, 1629-1630, 1631-1632, 1633-1634, 1635-1636, 1637-1638, 1639-1640, 1641-1642, 1643-1644, 1645-1646, 1647-1648, 1649-1650, 1651-1652, 1653-1654, 1655-1656, 1657-1658, 1659-1660, 1661-1662, 1663-1664, 1665-1666, 1667-1668, 1669-1670, 1671-1672, 1673-1674, 1675-1676, 1677-1678, 1679-1680, 1681-1682, 1683-1684, 1685-1686, 1687-1688, 1689-1690, 1691-1692, 1693-1694, 1695-1696, 1697-1698, 1699-1700, 1701-1702, 1703-1704, 1705-1706, 1707-1708, 1709-1710, 1711-1712, 1713-1714, 1715-1716, 1717-1718, 1719-1720, 1721-1722, 1723-1724, 1725-1726, 1727-1728, 1729-1730, 1731-1732, 1733-1734, 1735-1736, 1737-1738, 1739-1740, 1741-1742, 1743-1744, 1745-1746, 1747-1748, 1749-1750, 1751-1752, 1753-1754, 1755-1756, 1757-1758, 1759-1760, 1761-1762, 1763-1764, 1765-1766, 1767-1768, 1769-1770, 1771-1772, 1773-1774, 1775-1776, 1777-1778, 1779-1780, 1781-1782, 1783-1784, 1785-1786, 1787-1788, 1789-1790, 1791-1792, 1793-1794, 1795-1796, 1797-1798, 1799-1800, 1801-1802, 1803-1804, 1805-1806, 1807-1808, 1809-1810, 1811-1812, 1813-1814, 1815-1816, 1817-1818, 1819-1820, 1821-1822, 1823-1824, 1825-1826, 1827-1828, 1829-1830, 1831-1832, 1833-1834, 1835-1836, 1837-1838, 1839-1840, 1841-1842, 1843-1844, 1845-1846, 1847-1848, 1849-1850, 1851-1852, 1853-1854, 1855-1856, 1857-1858, 1859-1860, 1861-1862, 1863-1864, 1865-1866, 1867-1868, 1869-1870, 1871-1872, 1873-1874, 1875-1876, 1877-1878, 1879-1880, 1881-1882, 1883-1884, 1885-1886, 1887-1888, 1889-1890, 1891-1892, 1893-1894, 1895-1896, 1897-1898, 1899-1900, 1901-1902, 1903-1904, 1905-1906, 1907-1908, 1909-1910, 1911-1912, 1913-1914, 1915-1916, 1917-1918, 1919-1920, 1921-1922, 1923-1924, 1925-1926, 1927-1928, 1929-1930, 1931-1932, 1933-1934, 1935-1936, 1937-1938, 1939-1940, 1941-1942, 1943-1944, 1945-1946, 1947-1948, 1949-1950, 1951-1952, 1953-1954, 1955-1956, 1957-1958, 1959-1960, 1961-1962, 1963-1964, 1965-1966, 1967-1968, 1969-1970, 1971-1972, 1973-1974, 1975-1976, 1977-1978, 1979-1980, 1981-1982, 1983-1984, 1985-1986, 1987-1988, 1989-1990, 1991-1992, 1993-1994, 1995-1996, 1997-1998, 1999-2000, 2001-2002, 2003-2004, 2005-2006, 2007-2008, 2009-2010, 2011-2012, 2013-2014, 2015-2016, 2017-2018, 2019-2020, 2021-2022, 2023-2024, 2025-2026, 2027-2028, 2029-2030, 2031-2032, 2033-2034, 2035-2036, 2037-2038, 2039-2040, 2041-2042, 2043-2044, 2045-2046, 2047-2048, 2049-2050, 2051-2052, 2053-2054, 2055-2056, 2057-2058, 2059-2060, 2061-2062, 2063-2064, 2065-2066, 2067-2068, 2069-2070, 2071-2072, 2073-2074, 2075-2076, 2077-2078, 2079-2080, 2081-2082, 2083-2084, 2085-2086, 2087-2088, 2089-2090, 2091-2092, 2093-2094, 2095-2096, 2097-2098, 2099-2100, 2101-2102, 2103-2104, 2105-2106, 2107-2108, 2109-2110, 2111-2112, 2113-2114, 2115-2116, 2117-2118, 2119-2120, 2121-2122, 2123-2124, 2125-2126, 2127-2128, 2129-2130, 2131-2132, 2133-2134, 2135-2136, 2137-2138, 2139-2140, 2141-2142, 2143-2144, 2145-2146, 2147-2148, 2149-2150, 2151-2152, 2153-2154, 2155-2156, 2157-2158, 2159-2160, 2161-2162, 2163-2164, 2165-2166, 2167-2168, 2169-2170, 2171-2172, 2173-2174, 2175-2176, 2177-2178, 2179-2180, 2181-2182, 2183-2184, 2185-2186, 2187-2188, 2189-2190, 2191-2192, 2193-2194, 2195-2196, 2197-2198, 2199-2200, 2201-2202, 2203-2204, 2205-2206, 2207-2208, 2209-2210, 2211-2212, 2213-2214, 2215-2216, 2217-2218, 2219-2220, 2221-2222, 2223-2224, 2225-2226, 2227-2228, 2229-2230, 2231-2232, 2233-2234, 2235-2236, 2237-2238, 2239-2240, 2241-2242, 2243-2244, 2245-2246, 2247-2248, 2249-2250, 2251-2252, 2253-2254, 2255-2256, 2257-2258, 2259-2260, 2261-2262, 2263-2264, 2265-2266, 2267-2268, 2269-2270, 2271-2272, 2273-2274, 2275-2276, 2277-2278, 2279-2280, 2281-2282, 2283-2284, 2285-2286, 2287-2288, 2289-2290, 2291-2292, 2293-2294, 2295-2296, 2297-2298, 2299-2300, 2301-2302, 2303-2304, 2305-2306, 2307-2308, 2309-2310, 2311-2312, 2313-2314, 2315-2316, 2317-2318, 2319-2320, 2321-2322, 2323-2324, 2325-2326, 2327-2328, 2329-2330, 2331-2332, 2333-2334, 2335-2336, 2337-2338, 2339-2340, 2341-2342, 2343-2344, 2345-2346, 2347-2348, 2349-2350, 2351-2352, 2353-2354, 2355-2356, 2357-2358, 2359-2360, 2361-2362, 2363-2364, 2365-2366, 2367-2368, 2369-2370, 2371-2372, 2373-2374, 2375-2376, 2377-2378, 2379-2380, 2381-2382, 2383-2384, 2385-2386, 2387-2388, 238



One point not obvious is that  $R_2$  and  $R_3$  with  $R_1$  in parallel are actually in series with the input to the filter. The values chosen normally

The modulator is now ready to be connected to the succeeding stages of the exciter.



## CHARACTERISTICS OF SIDEBAND FILTER AS ACTUALLY CONSTRUCTED

The filter, for which the characteristics are shown on the next page, was constructed to the specifications in the article on filter design, with the exception that inductances of a much lower "Q" were substituted. The results of this substitution is readily apparent by comparing the solid curve below, with the corresponding characteristic curve of the designed filter. The attenuation peak at 20.5 KC is suppressed, and the attenuation in general is less over the suppressed sideband range. The most important effect, however, is the lowering of sharpness of cut-off from 20 to 21 KC.

In order that this filter be usable, the attenuation curve was shifted somewhat, as shown by the dashed curve, by shunting a .0017 mfd mica condenser across  $C_4$ .

The indication of an apparently negative attenuation in the bandpass region is due to a slight miss-match of impedance in the test circuit.









# Wideband

Design criteria are presented for L-C and R-C phase shifting networks that develop a constant phase difference over a wide frequency band. The technique is to introduce two phase shifts such that, although they vary with frequency, their difference does not

IT is frequently desirable to derive from a given voltage two new voltages of the same frequency but with the phase angle between the new voltages held substantially constant over a wide frequency range, each derived voltage having an amplitude characteristic linearly variable with the amplitude of the input voltage independently of frequency.

To produce the phase difference between the two voltages, two networks whose phase angles increase substantially linearly with the logarithm of the frequency are used. Thus, if the two networks are properly matched, the phase difference between them remains nearly con-

stant over a wide range of frequency.

## Network Development

One way of producing the two voltages is to derive both of them from a single source and arrange that the output of either channel is independent of frequency, but that both have phase angles with respect to the input voltages varying in such a way that over a wide band of frequencies

where  $\phi_1 = \phi_2 = K$   
 $\phi_1$ : phase angle of output No. 1  
 $\phi_2$ : phase angle of output No. 2  
 $K$ : a constant  
 $\phi_1$  and  $\phi_2$  are each their own function of  $f$ , the frequency

One possible configuration for these functions is

$$\phi_1 = C + \log f \quad 2$$

$$\phi_2 = C + \log kf \quad 3$$

where  $C$  and  $k$  are constants

Substituting Eq. 2 and 3 in Eq. 1

$$\begin{aligned} \phi_1 - \phi_2 &= C + \log f - C - \log kf \\ &= \log f - \log f - \log k \\ &= -\log k \equiv K \end{aligned} \quad 4$$

Thus it is only necessary to find a network configuration which will yield a phase angle which varies as the logarithm of the frequency over a wide range of frequency. The network must also have no change in output amplitude with frequency. The latter limitation usually restricts the final network configuration to lattice types because finite ladder types with any phase shift must be accompanied by amplitude variations. In order to avoid the use of transformers in the lattice structure, a half-lattice will be chosen with the input terminals excited by two equal voltages 180 degrees out of phase. Such voltages are readily available from vacuum tube phase inverters consisting of a tube with equal cathode and anode loads.

## L-C Lattice

A circuit having enough independent parameters to permit the designer to shape the phase angle curve to the required logarithmic form and the basic design equations of the circuit are shown in Fig. 1A. It will be noted from the equation for the phase angle (Fig. 1A) that an arbitrary factor  $s$  is included. The choice of  $s$  is up to the designer.

If  $s$  lies between 3 and 5, a fairly straight line for  $\phi$  is obtained when plotted against a logarithmic frequency scale. As will be shown later, if a second curve for a second similar network, but with a resonant frequency  $f_0$  which is 4.53 times the resonant frequency of the first curve, is plotted, the phase

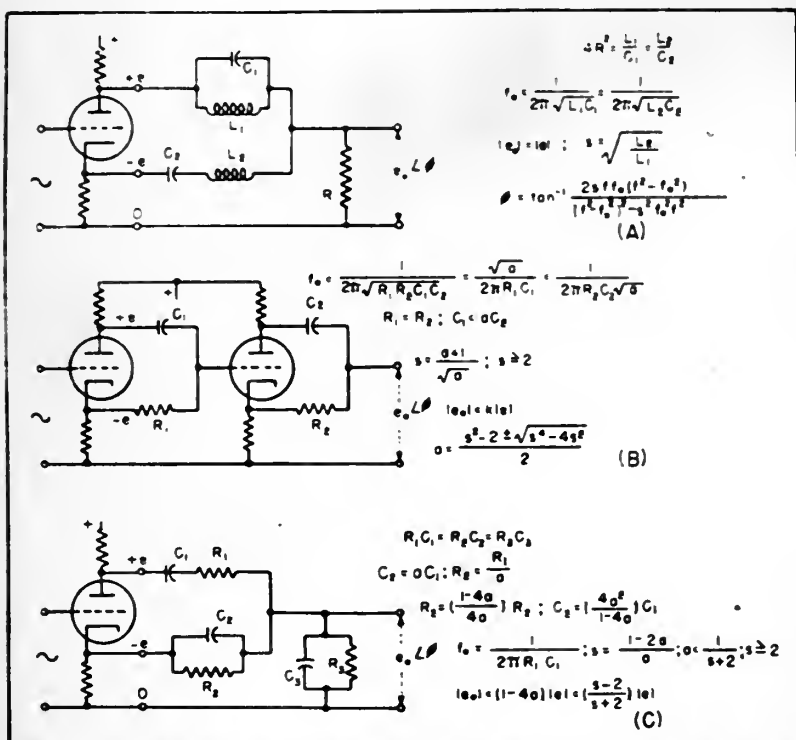


FIG. 1—Three networks for producing phase shift. Networks are used in pairs to obtain two outputs which have a constant phase difference over a wide range of frequency. Phase shift equation given for network of Fig. 1A holds for other two networks



# Phase Shift Networks

By R. B. DOME

Research Engineer,  
General Electric  
Bridgeport, Ct.

angle difference between these two curves is maintained within  $90 \pm 4$  degrees over a fairly wide range of frequency if  $s = 4$ , and is quite satisfactory for voice frequencies because the upper to lower frequency ratio is approximately 28 to 1, as for example would occur in a band from 130 to 3,600 cps.

The band can be widened indefinitely by adding more elements to the bridge arms with the result that the frequency range is expanded more until remarkably good performance is obtained over the full audio range from 30 to 15,000 cps.

The use of inductances in filters is to be avoided if possible because of the inability to obtain a pure inductance; there are always the inevitable series resistance and shunt capacitance. Other objections to the use of inductors are that they may pick up unwanted signals from stray magnetic fields and that they are not constant in inductance with applied voltage. As a consequence, a resistance-capacitance type of half-lattice network would be preferable providing the required type of phase curve were obtainable from them.

## R-C Lattice

In order to provide a wide range of operation, two simple resistance-capacitance networks can be used in cascade with isolating tubes between them as shown in Fig. 1B. A single network has restricted bandwidth. No terminating resistor can be used in this type of network.

The phase angle of the final output voltage can be adjusted exactly as in the L-C circuit of Fig. 1A because the phase angle equation is identical. The only limitation with this network is that  $s$  must be greater than 2, however as previously found, for best results  $s$

## Applications

THE wideband phase shift networks described herein are applicable to:

Single sideband telephony accomplished directly at the final carrier frequency without multiple modulators or sharp cutoff filters.

High efficiency broadcast transmitters radiating carrier, upper, and lower sidebands from three separate antennas.

Frequency variation of crystal controlled carrier of communication transmitters either for carrier control or frequency-shift keying.

Circular display on cathode-ray tubes over a wide frequency band without adjustment of the phase shift network.

Variable speed operation of three-phase a-c motors.

should be approximately 4, and hence has well above 2.

The number of stages and delay in cascade may, of course, be increased to any desired number. When six stages are used, the required delay for a phase angle difference between two sets of antennas can be maintained within one-half a degree because one stage gives 200 to 1, or greater, delay over an audio frequency band from 30 to 10,000 cps, which is adequate for high quality broadcast systems.

Another R-C network which does not require an isolating tube, is shown in Fig. 1C. This network somewhat resembles the L-C network of Fig. 1A but differs in output voltage. The output voltage is a proper fraction of the input voltage. The factor must be less than 0.25 or  $s$  must be greater than 2, when using this network. As previously pointed out, for best results  $s$  should be approximately 4, or  $\alpha = 0.167$  which will yield an output voltage of about 0.32 times the input voltage.

The phase angle equation (Fig. 1A) is seen to be identical to those

for the previously described networks. In order to find the proper values for  $t$  in the two networks which are to yield a phase angle difference of 90 degrees, assume a geometric mean frequency between the upper and lower frequency limits of usable phase angle difference. This mean frequency is not critical, but a value of 700 cps has been suggested as a practical frequency.

O. B. Hansen, Down to Earth on 'High Fidelity', Radio Technical Planning Board, Report on Standards and Frequency Allocations for Post War FM Broadcasting, Panel 5, June 1, 1944, Section VIII. Having decided on the mean frequency of the system, it is merely necessary to design the two phase shift networks so that one has a phase angle of  $180 - 45$  degrees at 700 cps, and the other a phase angle of  $180 + 45$  degrees at 700 cps.

## Basis for Design

Suppose the value for  $t$  of the first network is to be determined, the  $t_0$  designating  $t$  of the first network as  $t_0$  and letting  $f$  equal the mean frequency, designated as  $F$ , the phase angle equation (Fig. 1) becomes

$$\tan \alpha = \frac{2F}{t_0} \sqrt{\frac{F}{t_0}} = \frac{2F^2}{t_0^2}$$

Setting  $\alpha = 135$  degrees and  $s = 4$ , clearing fractions and solving for  $t_0$  gives  $t_0 = 2.126F$ , which, for  $F = 700$  cps, gives  $t_0 = 1,488$  cps.

By reciprocal relationships, the value of  $t$ , the  $t_1$  of the second network, is  $t_1 = F/2.126$ ; at  $F = 700$  cps,  $t_1 = 329$  cps.

The parameters of the two networks can now be calculated. For convenience and to establish values of impedance into which vacuum tubes can work satisfactorily, assume that  $R = 20,000$  ohms in each



Table I—Determination of Network Parameters

First Network	Second Network
$f_{01} = 1133 \text{ cps}$	$f_{02} = 329 \text{ cps}$
$s = 1.00$	$s = 1.00$
$a = 0.1666$	$a = 0.1666$
$R_1 = 20,000 \text{ ohms}$	$R_1 = 20,000 \text{ ohms}$
$C_1 = \frac{1}{2\pi f_{01} R_1} = \frac{1}{2\pi 1133 \times 20000}$ $= 0.00535 \text{ } \mu\text{f}$	$C_1 = \frac{1}{2\pi f_{02} R_1} = \frac{1}{2\pi 329 \times 20000}$ $= 0.0242 \text{ } \mu\text{f}$
$C_2 = a C_1 = 0.166 \times 0.00535$ $= 0.000892 \text{ } \mu\text{f}$	$C_2 = a C_1 = 0.166 \times 0.0242$ $= 0.00403 \text{ } \mu\text{f}$
$C = \left( \frac{1a}{1 - 1a} \right) C_1 = 0.333 C_1$ $= 0.333 \times 0.00535$ $= 0.001785 \text{ } \mu\text{f}$	$C = \left( \frac{1a}{1 - 1a} \right) C_1 = 0.333 C_1$ $= 0.333 \times 0.0242$ $= 0.00806 \text{ } \mu\text{f}$
$R = \frac{R_1}{a} = \frac{20000}{0.166}$ $= 120,000 \text{ ohms}$	$R = \frac{R_1}{a} = \frac{20000}{0.166}$ $= 120,000 \text{ ohms}$
$R_2 = \left( \frac{1 - 1a}{1a} \right) R$ $= \left( \frac{1 - 0.666}{0.666} \right) 120,000$ $= 60,000 \text{ ohms}$	$R_2 = \left( \frac{1 - 1a}{1a} \right) R$ $= \left( \frac{1 - 0.666}{0.666} \right) 120,000$ $= 60,000 \text{ ohms}$

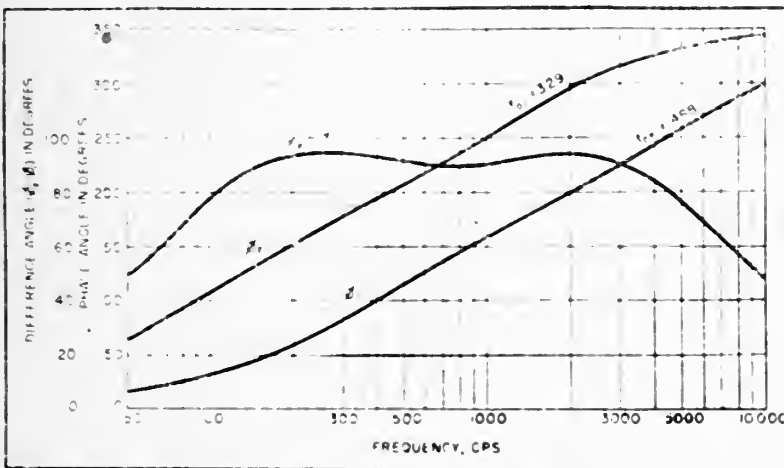


FIG. 2—Phase shift characteristics of networks whose values are given in Table I

network. The calculations are summarized in Table I.

The phase shift curves for these two networks are shown in Fig. 2. Note the long frequency range over which these curves are substantially parallel. The difference angle is plotted to facilitate comparison. Note that this difference angle holds fairly close to 90 degrees over a range of frequencies from 130 to 3,600 cps. This range is quite adequate for a voice frequency channel.

Such a network can be used to build a simple single sideband transmitter without the necessity of using sharp cut-off filters and double or triple modulation. This simplification may be done by combining the outputs of two balanced modulators: one balanced modulator is fed with radio frequency  $\cos \omega t$  and audio frequency  $m \cos (\mu t - \theta)$ , and the other balanced modulator is fed with radio frequency  $\cos (\omega t \pm \pi/2)$  and audio frequency  $m \cos (\mu t - \theta \pm \pi/2)$ . If

the plus signs are used and the balanced modulator outputs are added, the resultant output would be  $m \cos [(\omega - \mu)t + \theta]$ , which is the lower sideband.

#### Single Sideband Telephony

An experimental system of this type has been constructed and operated using carrier frequencies in the broadcast band. No untoward difficulty was encountered and results were fully up to expectations. The transmitter was modulated with audio signals taken from the output of a broadcast receiver. The single sideband transmission was received by a second broadcast receiver into whose input terminals was also fed an unmodulated continuous wave radio frequency from a signal generator to furnish the missing carrier. The signal generator frequency was adjusted until the reproduced program sounded most natural. No auxiliary filters were used in the radio section of the transmitter to aid in reducing the unwanted sideband.

The question may arise as to how much of the unwanted sideband is permitted to get through the system if not exactly 90 degrees phase difference is obtained between the two audio channels. The ratio of the weaker sideband to the stronger sideband is given by the equation

$$\text{RATIO} = \sqrt{\frac{1 - \cos \delta}{1 + \cos \delta}} \cong \frac{\delta}{2}$$

where  $\delta$  is the deviation of the difference phase angle from  $\pi/2$ , with  $\delta$  expressed in radians for the approximate formula. A deviation from 90 degrees of 6 degrees is required before the weaker sideband becomes 5 percent as strong as the stronger sideband, so that the phase angle curve of Fig. 2 is probably commercially useful over a range in frequencies included between 84 degrees on the low side of the center to 84 degrees on the high side of the center.

A second application for the system is that of providing a closer-to-zero frequency single sideband telephone or broadcast transmission when filters are used. For example, a single sideband transmitter may be constructed along conventional lines with a quartz crystal type of





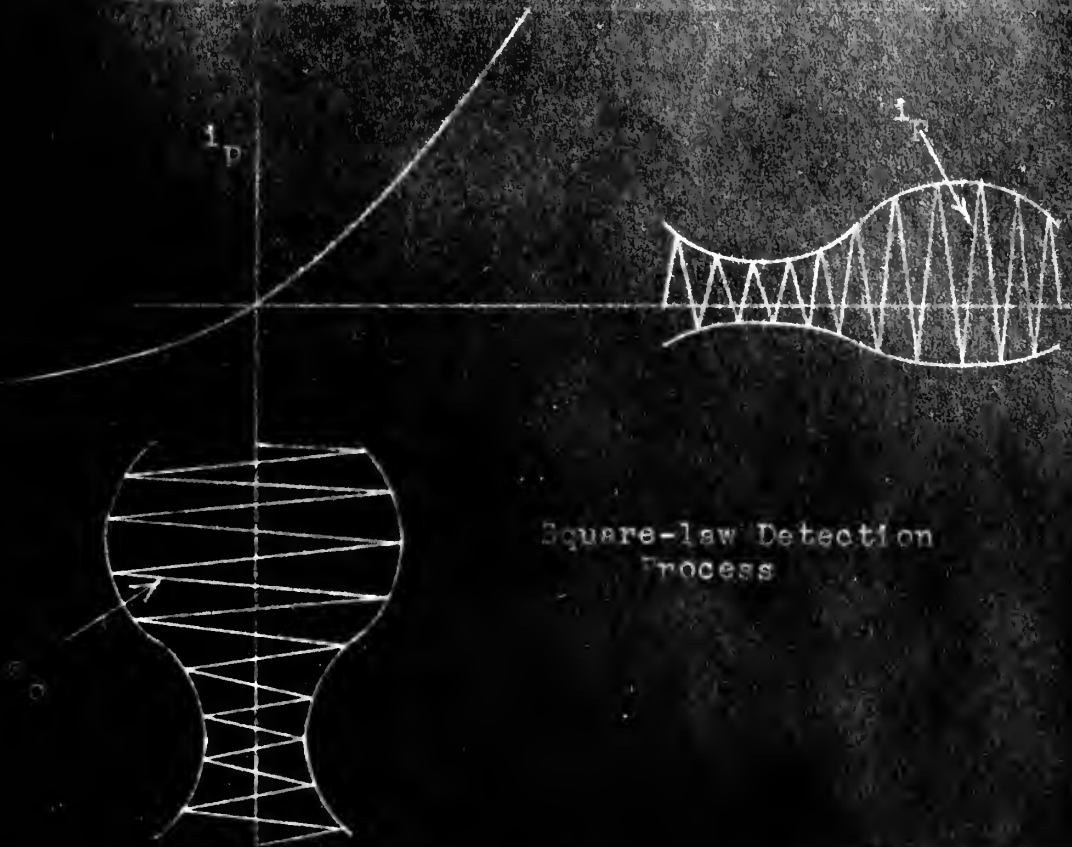




# APPENDIX III

## DERIVATION OF FREQUENCY COMPONENTS RESULTING FROM SQUARE LAW-DETECTION

### Case I. Detection of normal A.M.



$$e_o = E_c \cos \omega_c t + \frac{1}{2} E_c M \cos (\omega_c + \omega_s) t \\ + \frac{1}{2} E_c M \cos (\omega_c - \omega_s) t$$

$$i_p = a_1 e_o + a_2 e_o^2$$



$$\begin{aligned}
 i_p = a_1 \{ e_0 \} + a_2 E_c^2 \{ & \cos \omega_c t \\
 & + \frac{M^2}{4} \cos^2(\omega_c + \omega_s)t + \frac{M^2}{4} \cos^2(\omega_c - \omega_s)t \\
 & + M \cos \omega_c t \cos(\omega_c + \omega_s)t \\
 & + M \cos \omega_c t \cos(\omega_c - \omega_s)t \\
 & + \frac{M^2}{2} \cos(\omega_c + \omega_s)t \cos(\omega_c - \omega_s)t \}
 \end{aligned}$$

$$\begin{aligned}
 i_p = a_1 \{ e_0 \} + a_2 E_c^2 \{ & \frac{1}{2} [1 + \cos 2\omega_c t] \\
 & + \frac{M^2}{8} [1 + \cos 2(\omega_c + \omega_s)t] \\
 & + \frac{M^2}{8} [1 + \cos 2(\omega_c - \omega_s)t] \\
 & + \frac{M}{2} [\cos(\omega_c - \omega_c - \omega_s)t + \cos(\omega_c + \omega_c + \omega_s)t] \\
 & + \frac{M}{2} [\cos(\omega_c - \omega_c + \omega_s)t + \cos(\omega_c + \omega_c - \omega_s)t] \\
 & + \frac{M^2}{4} [\cos(\omega_c + \omega_s - \omega_c + \omega_s)t + \cos(\omega_c + \omega_s + \omega_c - \omega_s)t]
 \end{aligned}$$

$$\begin{aligned}
 i_p = a_1 \{ e_0 \} + a_2 E_c^2 \left( \frac{1}{2} + \frac{M^2}{8} + \frac{M^2}{8} \right) \\
 + a_2 E_c^2 \left\{ \frac{1}{2} \cos 2\omega_c t + \frac{M^2}{8} \cos 2(\omega_c + \omega_s)t \right. \\
 + \frac{M^2}{8} \cos 2(\omega_c - \omega_s)t + \frac{M}{2} \cos \omega_s t \\
 + \frac{M}{2} \cos(2\omega_c + \omega_s)t + \frac{M}{2} \cos \omega_s t \\
 + \frac{M}{2} \cos(2\omega_c - \omega_s)t + \frac{M^2}{4} \cos 2\omega_s t \\
 \left. + \frac{M^2}{4} \cos 2\omega_c t \right\}
 \end{aligned}$$



$$\begin{aligned}
 i_p = & a_2 E_c^2 \left( \frac{1}{2} + \frac{M^2}{4} \right) + a_1 E_c M \cos \omega_s t + a_2 E_c \cos \omega_c t \\
 & + a_2 E_c^2 \frac{M^2}{4} \cos 2\omega_s t + a_1 E_c \cos \omega_c t \\
 & + a_2 E_c^2 \left( \frac{1}{2} + \frac{M^2}{4} \right) \cos 2\omega_c t \\
 & + \frac{1}{2} a_1 E_c M \cos(\omega_c + \omega_s) t + \frac{1}{2} a_1 E_c M \cos(\omega_c - \omega_s) t \\
 & + a_2 E_c^2 \frac{M^2}{8} \cos 2(\omega_c + \omega_s) t + a_2 E_c^2 \frac{M^2}{8} \cos 2(\omega_c - \omega_s) t \\
 & + a_2 E_c^2 \frac{M}{2} \cos(2\omega_c + \omega_s) t + a_2 E_c^2 \frac{M}{2} \cos(2\omega_c - \omega_s) t
 \end{aligned}$$

Components of  $i_p$  :

- |   |  |
|---|--|
| (1) $a_2 E_c^2 \left( \frac{1}{2} + \frac{M^2}{4} \right)$                  | -- D. C. Component                                 |
| (2) $a_2 E_c^2 M \cos \omega_s t$   | -- Signal Frequency Component                      |
| (3) $a_2 E_c^2 \frac{M^2}{4} \cos 2\omega_s t$                              | -- Double Signal Frequency Component               |
| (4) $a_1 E_c \cos \omega_c t$   | -- Carrier Frequency Component                     |
| (5) $a_2 E_c^2 \left( \frac{1}{2} + \frac{M^2}{4} \right) \cos 2\omega_c t$ | -- Double Carrier Frequency Component              |
| (6) $\frac{1}{2} a_1 E_c M \cos(\omega_c + \omega_s) t$                     | -- Upper Sideband Component                        |
| (7) $\frac{1}{2} a_1 E_c M \cos(\omega_c - \omega_s) t$                     | -- Lower Sideband Component                        |
| (8) $a_2 E_c^2 \frac{M^2}{8} \cos 2(\omega_c + \omega_s) t$                 | -- Double Upper Sideband Frequency Component       |
| (9) $a_2 E_c^2 \frac{M^2}{8} \cos 2(\omega_c - \omega_s) t$                 | -- Double Lower Sideband Frequency Component       |
| (10) $a_2 E_c^2 \frac{M}{2} \cos(2\omega_c + \omega_s) t$                   | -- Double Carrier Plus Signal Frequency Component  |
| (11) $a_2 E_c^2 \frac{M}{2} \cos(2\omega_c - \omega_s) t$                   | -- Double Carrier Minus Signal Frequency Component |





Case II.  $\omega_c \neq \omega_s$  (detuning). The input signal is assumed to be a frequency of  $\omega_c + \Delta$ .

$$e_o = C \cos(\omega_c + \Delta)t + \frac{1}{2} E_c M \cos(\omega_c + \omega_s)t + \frac{1}{2} E_c M \cos(\omega_c - \omega_s)t$$

$$i_p = a_1 e_o + a_2 e_o^2$$

$$i_p = a_1 \{e_o\} + a_2 \left\{ C^2 \cos^2(\omega_c + \Delta)t + \frac{E_c^2 M^2}{4} \cos^2(\omega_c + \omega_s)t + \frac{E_c^2 M^2}{4} \cos^2(\omega_c - \omega_s)t + C E_c M \cos(\omega_c + \Delta)t \cos(\omega_c + \omega_s)t + C E_c M \cos(\omega_c + \Delta)t \cos(\omega_c - \omega_s)t + \frac{E_c^2 M^2}{2} \cos(\omega_c + \omega_s)t \cos(\omega_c - \omega_s)t \right\}$$

$$i_p = a_2 \left( \frac{C^2}{2} + \frac{E_c^2 M^2}{4} \right) + \frac{a_2 C E_c M}{2} \cos(\omega_s - \Delta)t + \frac{a_2 C E_c M}{2} \cos(\omega_s + \Delta)t + \frac{a_2 E_c^2 M^2}{4} \cos 2\omega_s t + a_1 C \cos(\omega_c + \Delta)t + a_2 \frac{E_c^2 M^2}{4} \cos 2\omega_c t + a_2 \frac{C^2}{2} \cos 2(\omega_c + \Delta)t + a_1 \frac{E_c M}{2} \cos(\omega_c + \omega_s)t + a_1 \frac{E_c M}{2} \cos(\omega_c - \omega_s)t + a_2 \frac{E_c^2 M^2}{8} \cos 2(\omega_c + \omega_s)t + a_2 \frac{E_c^2 M^2}{8} \cos 2(\omega_c - \omega_s)t + a_2 \frac{C E_c M}{2} \cos(2\omega_c + \Delta + \omega_s)t + a_2 \frac{C E_c M}{2} \cos(2\omega_c + \Delta - \omega_s)t$$





- (1)  $A_1 C \cos(\omega_c t)$
  - (2)  $\frac{a_1}{2} C E_c M \cos(\omega_c + \Delta)t$
  - (3)  $\frac{a_1}{2} C E_c M \cos(\omega_c - \Delta)t$
  - (4)  $\frac{a_1}{4} E_c^2 M^2 \cos 2\omega_c t$
  - (5)  $a_1 C \cos(\omega_c + \Delta)t$
  - (6)  $\frac{a_1}{4} E_c^2 M^2 \cos 2\omega_c t$
  - (7)  $\frac{a_1 C^2}{2} \cos 2(\omega_c + \Delta)t$
  - (8)  $\frac{a_1 E_c M}{2} \cos(\omega_c + \omega_a)t$
  - (9)  $\frac{a_1 E_c M}{2} \cos(\omega_c - \omega_a)t$
  - (10)  $\frac{a_1}{8} E_c^2 M^2 \cos 2(\omega_c + \omega_a)t$
  - (11)  $\frac{a_1}{8} E_c^2 M^2 \cos 2(\omega_c - \omega_a)t$
  - (12)  $\frac{a_1}{2} C E_c M \cos(2\omega_c + \Delta + \omega_a)t$
  - (13)  $\frac{a_1}{2} C E_c M \cos(2\omega_c + \Delta - \omega_a)t$
- Double Frequency  
-- Double Frequency  
-- Double Frequency  
-- Double Frequency  
-- Double Frequency  
-- Double Frequency  
-- Double Frequency  
-- Upper Sideband  
-- Lower Sideband  
-- Double Upper Sideband Frequency  
-- Double Lower Sideband Frequency  
-- Reinserted Carrier Plus Upper Sideband Frequency  
-- Reinserted Carrier Plus Lower Sideband Frequency

Note that components (2) and (3) may be combined to give  $a_1 C E_c M \cos \Delta t \cos \omega_c t$  which is a signal similar to the audio signal of straight A.M., detected, but varying in amplitude from zero to  $a_1 C E_c M$  at the rate  $\Delta$ . The result would be unintelligible.



Case 1:  $\omega_c \neq \omega_s$  (The carrier and signal frequencies are not equal)  
 In this case, the output current is given by  
 $i_p = a_1 e_o + a_2 e_o^2$

$$e_o = C \cos(\omega_c t + \theta) + \frac{E_c M}{2} \cos(\omega_c + \omega_s)t + \frac{E_c M}{2} \cos(\omega_c - \omega_s)t$$

$$i_p = a_1 e_o + a_2 e_o^2$$

$$\begin{aligned} i_p = a_1 \{ e_o \} + a_2 \{ & C^2 \cos^2(\omega_c t + \theta) \\ & + \frac{E_c^2 M^2}{4} \cos^2(\omega_c + \omega_s)t + \frac{E_c^2 M^2}{4} \cos^2(\omega_c - \omega_s)t \\ & + C E_c M \cos(\omega_c t + \theta) \cos(\omega_c + \omega_s)t \\ & + C E_c M \cos(\omega_c t + \theta) \cos(\omega_c - \omega_s)t \\ & + \frac{E_c^2 M^2}{2} \cos(\omega_c + \omega_s)t \cos(\omega_c - \omega_s)t \} \end{aligned}$$

$$\begin{aligned} i_p = a_1 \left( \frac{C^2}{2} + \frac{E_c^2 M^2}{4} \right) + \frac{a_1 C E_c M}{2} \cos(\omega_s t - \theta) \\ + \frac{a_1 C E_c M}{2} \cos(\omega_s t + \theta) + \frac{a_1 E_c^2 M^2}{4} \cos 2\omega_s t \\ + a_1 C \cos(\omega_c t + \theta) + \frac{a_1 E_c M}{2} \cos(\omega_c + \omega_s)t \\ + \frac{a_1 E_c M}{2} \cos(\omega_c - \omega_s)t + \frac{a_1 C^2}{2} \cos 2(\omega_c t + \theta) \\ + \frac{a_1 E_c^2 M^2}{8} \cos 2(\omega_c + \omega_s)t + \frac{a_1 E_c^2 M^2}{8} \cos 2(\omega_c - \omega_s)t \\ + \frac{a_1 C E_c M}{2} \cos(2\omega_c t + \theta + \omega_s t) \\ + \frac{a_1 C E_c M}{2} \cos(2\omega_c t + \theta - \omega_s t) + \frac{a_1 E_c^2 M^2}{4} \cos 2\omega_c t \end{aligned}$$



- (1)  $a_1 \left( \frac{C}{2} + \frac{M}{2} \right)$  -- Carrier Frequency
- (2)  $\frac{a_2}{2} C E_c M \cos(\omega_c t + \theta)$  -- Reinserted Carrier
- (3)  $\frac{a_2}{2} C E_c M \cos(\omega_c t + \theta)$  -- Reinserted Carrier
- (4)  $\frac{a_2}{4} E_c^2 M^2 \cos 2 \omega_s t$  -- Double Signal Frequency
- (5)  $a_1 C \cos(\omega_c t + \theta)$  -- Reinserted Carrier
- (6)  $\frac{a_2}{4} E_c^2 M^2 \cos 2 \omega_c t$  -- Double Carrier Frequency
- (7)  $\frac{a_2}{2} \frac{C^2}{2} \cos 2(\omega_c t + \theta)$  -- Double Reinserted Carrier Frequency
- (8)  $\frac{a_1}{2} E_c M \cos(\omega_c + \omega_s)t$  -- Upper Sideband
- (9)  $\frac{a_1}{2} E_c M \cos(\omega_c - \omega_s)t$  -- Lower Sideband
- (10)  $\frac{a_2}{8} E_c^2 M^2 \cos 2(\omega_c + \omega_s)t$  -- Double Upper Sideband Frequency
- (11)  $\frac{a_2}{8} E_c^2 M^2 \cos 2(\omega_c - \omega_s)t$  -- Double Lower Sideband Frequency
- (12)  $\frac{a_2}{2} C E_c M \cos(2\omega_c t + \theta + \omega_s t)$  -- Reinserted Carrier Plus Upper Sideband Frequency
- (13)  $\frac{a_2}{2} C E_c M \cos(2\omega_c t + \theta - \omega_s t)$  -- Reinserted Carrier Plus Lower Sideband Frequency

Note that components (2) and (3) may be combined to give

$$a_2 C E_c M \cos \theta \cos \omega_s t,$$

which is equal to signal component of Case 1, but

multiplied by  $\cos \theta$ , which for  $\theta \neq 0$ , will reduce the value of this component.



$$e_s = \frac{E_c M}{2} \cos(\omega_c + \omega_s)t + \frac{E_c M}{2} \cos(\omega_c - \omega_s)t$$

$$i_p = a_1 e_s + a_2 e_s^2$$

$$i_p = a_1 \{e_s\} + a_2 \frac{E_c^2 M^2}{4} \{ \cos^2(\omega_c + \omega_s)t + \cos^2(\omega_c - \omega_s)t + 2 \cos(\omega_c + \omega_s)t \cos(\omega_c - \omega_s)t \}$$

$$i_p = \frac{a_2 E_c^2 M^2}{4} + \frac{a_2 E_c^2 M^2}{4} \cos 2\omega_s t + \frac{a_1 E_c M}{2} \cos(\omega_c + \omega_s)t + \frac{a_1 E_c M}{2} \cos(\omega_c - \omega_s)t + \frac{a_2 E_c^2 M^2}{8} \cos 2(\omega_c + \omega_s)t + \frac{a_2 E_c^2 M^2}{8} \cos 2(\omega_c - \omega_s)t + \frac{a_2 E_c^2 M^2}{4} \cos 2\omega_c t$$

Components of  $i_p$ :

- |  |                                    |
|--|------------------------------------|
| (1) $\frac{a_2 E_c^2 M^2}{4}$                              | -- D. C. Component                 |
| (2) $\frac{a_2 E_c^2 M^2}{4} \cos 2\omega_s t$             | -- Double Signal Frequency         |
| (3) $\frac{a_2 E_c^2 M^2}{4} \cos 2\omega_c t$             | -- Double Carrier Frequency        |
| (4) $\frac{a_1 E_c M}{2} \cos(\omega_c + \omega_s)t$       | -- Upper Sideband                  |
| (5) $\frac{a_1 E_c M}{2} \cos(\omega_c - \omega_s)t$       | -- Lower Sideband                  |
| (6) $\frac{a_2 E_c^2 M^2}{8} \cos 2(\omega_c + \omega_s)t$ | -- Double Upper Sideband Frequency |
| (7) $\frac{a_2 E_c^2 M^2}{8} \cos 2(\omega_c - \omega_s)t$ | -- Double Lower Sideband Frequency |





Case 1:

$$e_o = \frac{E_c M}{2} \cos(\omega_c + \omega_s)t$$

$$i_p = a_1 e_o + a_2 e_o^2$$

$$i_p = \frac{a_1 E_c M}{2} \cos(\omega_c + \omega_s)t + \frac{a_2 E_c^2 M^2}{8} \cos^2(\omega_c + \omega_s)t$$

$$i_p = \frac{a_2 E_c^2 M^2}{8} + \frac{a_1 E_c M}{2} \cos(\omega_c + \omega_s)t + \frac{a_2 E_c^2 M^2}{8} \cos 2(\omega_c + \omega_s)t$$

Components of  $i_p$ :

- (1)  $\frac{a_2 E_c^2 M^2}{8}$  -- D.C. Component
- (2)  $\frac{a_1 E_c M}{2} \cos(\omega_c + \omega_s)t$  -- Upper Sideband Frequency
- (3)  $\frac{a_2 E_c^2 M^2}{8} \cos 2(\omega_c + \omega_s)t$  -- Lower Sideband Frequency

No signal frequency component is present, thus detection is not possible.



$$= C \cos(\omega_1 t + \phi) \cos(\omega_2 t + \phi)$$

$$= C \cos \phi \cos \omega_1 t + C \sin \phi \sin \omega_1 t$$

$$= \frac{E M}{\sqrt{2} \sqrt{1 + \cos 2\phi}}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$

$$= \frac{E M}{2}$$













(1)

(2)

Note that the ratio of the above components  
which are the same value obtained in appendix  
... by example, to be a  
... determining the relative  
... components resulting







## APPENDIX V. Hetrodyne Detection Analysis

(From "Theory and Applications of Electron Tubes  
-- H. J. Reich.)

5-14. Ideal Linear Diode Detection of the Sum of Two Voltages of Different Frequencies. The linear diode detector is sometimes used to produce a difference frequency as the result of the simultaneous application of two sinusoidal voltages of

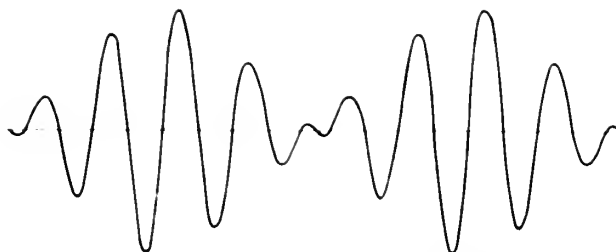


FIG. 5-14. Resultant of two sine waves of equal amplitude and 5 to 4 frequency ratio.

unequal frequency. To analyze this process it is necessary first to study the form of the resultant wave produced by the addition of two sine waves. Figure 5-14 shows the resultant of two sine waves of equal amplitude and 5 to 4 frequency ratio. By adding waves that have both unequal frequency and unequal amplitude, the reader may show that, in general, the resultant of two sine waves of somewhat different frequency resembles an amplitude-modulated wave. It differs from such a wave in that the points at which the resultant crosses the axis are not equally spaced, indicating additional phase modulation.

The form of the envelope of the resultant wave and the manner in which the phase varies with time can be determined analytically. Let the two impressed voltages be  $e_1 = E_1 \cos \omega_1 t$  and  $e_2 = E_2 \cos \omega_2 t$ . The resultant voltage is

$$e = E_1 \cos \omega_1 t + E_2 \cos \omega_2 t \quad (5-35)$$

$$= E_1 \cos \omega_1 t + E_2 \cos [(\omega_1 - \omega_2)t] \quad (5-36)$$

<sup>1</sup> RÖDER, HANS, *Proc. I. R. E.*, **20**, 1946 (1932).



By substituting  $\phi$  from (5-36) and collecting terms in  $\cos(\omega_1 - \omega_2)t$ , we obtain, formed into

$$E = E_1 \sqrt{1 + 2h \cos(\omega_1 - \omega_2)t + h^2} \quad (5-37)$$

in which

$$E = E_1 \sqrt{1 + 2h \cos(\omega_1 - \omega_2)t + h^2} \quad (5-38)$$

$$\phi = \tan^{-1} \frac{h \sin(\omega_1 - \omega_2)t}{1 + h \cos(\omega_1 - \omega_2)t} \quad (5-39)$$

$$h = \frac{E_2}{E_1} \quad (5-40)$$

If  $e_1$  represents the component of the applied voltage that has the larger amplitude, then  $E_1 \geq E_2$ , and  $h \leq 1$ .

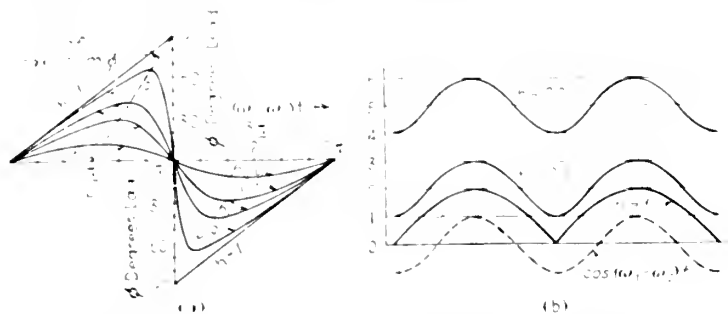


FIG. 5-15—Effect of amplitude ratio  $h$  upon (a) phase modulation and (b) form of envelope of the resultant of two sine waves of unequal frequency.

Figure 5-15a shows how the phase angle  $\phi$  varies throughout one cycle of difference frequency,  $f_1 - f_2$ , for various values of  $h$  between 0.2 and 1. When the amplitudes are equal,  $h = 1$ , and  $\phi$  varies between  $\pm 90^\circ$ ; a sudden reversal of phase takes place at the instant  $(\omega_1 - \omega_2)t = \pi$ . As  $h$  is decreased by changing the amplitude of either component, the maximum value of  $\phi$  rapidly decreases, being only slightly greater than  $11\frac{1}{2}^\circ$  when  $h = 0.2$ . Figure 5-15b shows how  $E$  varies throughout the difference-frequency cycle, for various values of  $h$ ,  $E_2$  being kept constant at unity. Comparison with the dotted curve, which represents  $\cos(\omega_1 - \omega_2)t$ , shows that the amplitude varies at a fundamental frequency equal to the difference frequency but also contains a steady component and harmonics of the difference frequency. E. B. Moullin has shown<sup>1</sup> that

<sup>1</sup> MOULLIN, E. B., *Wireless Eng.*, **9**, 378 (1932). See also F. M. Colebrook, *Wireless Eng.*, **9**, 195 (1932).





$$\begin{aligned}
& \sqrt{1 - 2h \cos \omega_1 t + h^2} = \sum_{n=0}^{\infty} P_n(\cos \omega_1 t) h^n \\
& \left( 1 + \frac{h^2}{2} + \frac{h^4}{2} + \frac{h^6}{2} + \dots \right) + j \left( 1 - \frac{h^2}{2} + \frac{h^4}{2} - \frac{5h^6}{2} + \dots \right) \cos 2\omega_1 t + \\
& \cos \omega_1 t + \omega_1 t - \frac{h}{4} \left( 1 - \frac{h^2}{4} - \frac{5h^4}{2} + \dots \right) \cos 2\omega_1 t + \\
& \omega_1 t + \frac{h}{8} \left( 1 - \frac{5h^2}{2} - \frac{7h^4}{2} + \dots \right) \cos 3\omega_1 t + \omega_1 t - \frac{5h^3}{2} \left( 1 - \frac{7h^2}{20} - \frac{21h^4}{5 \times 2^3} + \dots \right) \cos 4\omega_1 t + \dots \quad (5-41)
\end{aligned}$$

Each coefficient of this series is in itself an infinite series which converges rapidly if  $h \ll 1$ . Considering only the terms given in Eq. (5-41) the series is accurate to within 1 per cent even when  $h = 1$ . Equation (5-41) proves rigorously that  $E$  contains a steady component, a fundamental difference-frequency component, and all harmonics of the difference frequency.

This analysis shows that the resultant of two sine waves of different frequencies is equivalent to a wave whose frequency is that of the component of higher amplitude and whose amplitude and phase are modulated simultaneously. The signal voltage with which the amplitude of this equivalent wave is modulated consists of the fundamental difference frequency and all its harmonics. The phase modulation, the fundamental frequency of which is also equal to the difference frequency, may be made as small as desired by making  $h$  small. Under the assumption that  $h$  is small enough so that the phase modulation is negligible, the sum of the two sine waves is equivalent to an amplitude-modulated wave  $E_1 f(t) \cos \omega_1 t$ , in which  $f(t)$  is given by Eq. (5-41). In accordance with the principle explained in Sec. 5-13, the application of this voltage to a linear diode detector will result in an output voltage of the form  $DE_1 f(t)$ , where  $D$  is the detection efficiency given by the curves of Fig. 5-12.

An examination of the coefficients of Eq. 5-41 shows that the ratio of the amplitude of the  $n$ th harmonic to the amplitude of the fundamental difference-frequency output of the detector is roughly proportional to  $h^{n-1}$ .<sup>1</sup> Therefore the harmonic content may be made as small as desired by making the amplitude of one

<sup>1</sup> The exactness of this proportionality increases rapidly as  $h$  becomes smaller.



one of the input voltages larger than the other. The amplitude of the fundamental component of detector output voltage is

$$Fund. E = E_1 D \left( 1 - \frac{h^2}{2} - \frac{h^2}{2} + \frac{5h^2}{2} \right) \quad (5-42)$$

As  $h$  is decreased, this approaches the limiting value

$$Fund. E = E_1 D h = E_1 D \quad (5-43)$$

This shows that, if one input voltage is considerably larger than the other, the amplitude of the fundamental difference-frequency component of linear diode detector output depends only upon the amplitude of the smaller input voltage.<sup>1</sup> This is of importance in the design of superheterodyne radio receivers and heterodyne oscillators (see page 351).

**5-15. Effect of Curvature of Diode Characteristic.** Equations (5-23) and (5-30) hold only for linear characteristics. If the diode characteristic is not linear, the wave of plate current is not an exact replica of the upper half of the applied modulated voltage. This fact suggests that the output contains harmonics of the signal voltage. In general, the relation between total instantaneous diode plate current and plate voltage may be expressed in terms of a series



## APPENDIX VI. Construction of a SSSC Adaptor

The adaptor, whose construction is to be described, was designed by the General Electric Company.<sup>2</sup> It was designed to be used with a good quality communications receiver, making possible the reception of SSSC transmissions, the reception of only one sideband (either) of an A.M. transmission, the reception of narrow-band FM transmissions, and the reception of PM transmissions.

As may be seen from the circuit diagram, this adaptor follows the phasing method of sideband separation, together with reinserted (or exalted) carrier and reactance-tube lock-in circuit. A type 6AK6 tube is employed to provide proper impedance match between the adaptor and the receiver I.F. output. An audio stage and a voltage-regulated, highly filtered power supply complete the circuit. Positions 1 and 2 of Switch ABC permit reception on either upper or lower sideband. Position 3 permits reception of both sidebands with the local oscillator giving exalted carrier reception. Position 4 is for normal reception.

The physical lay-out of the components closely followed the circuit diagram arrangement. The 6AK6, with it's associated components was mounted in a small shield can, and located in the receiver chassis. The small filament



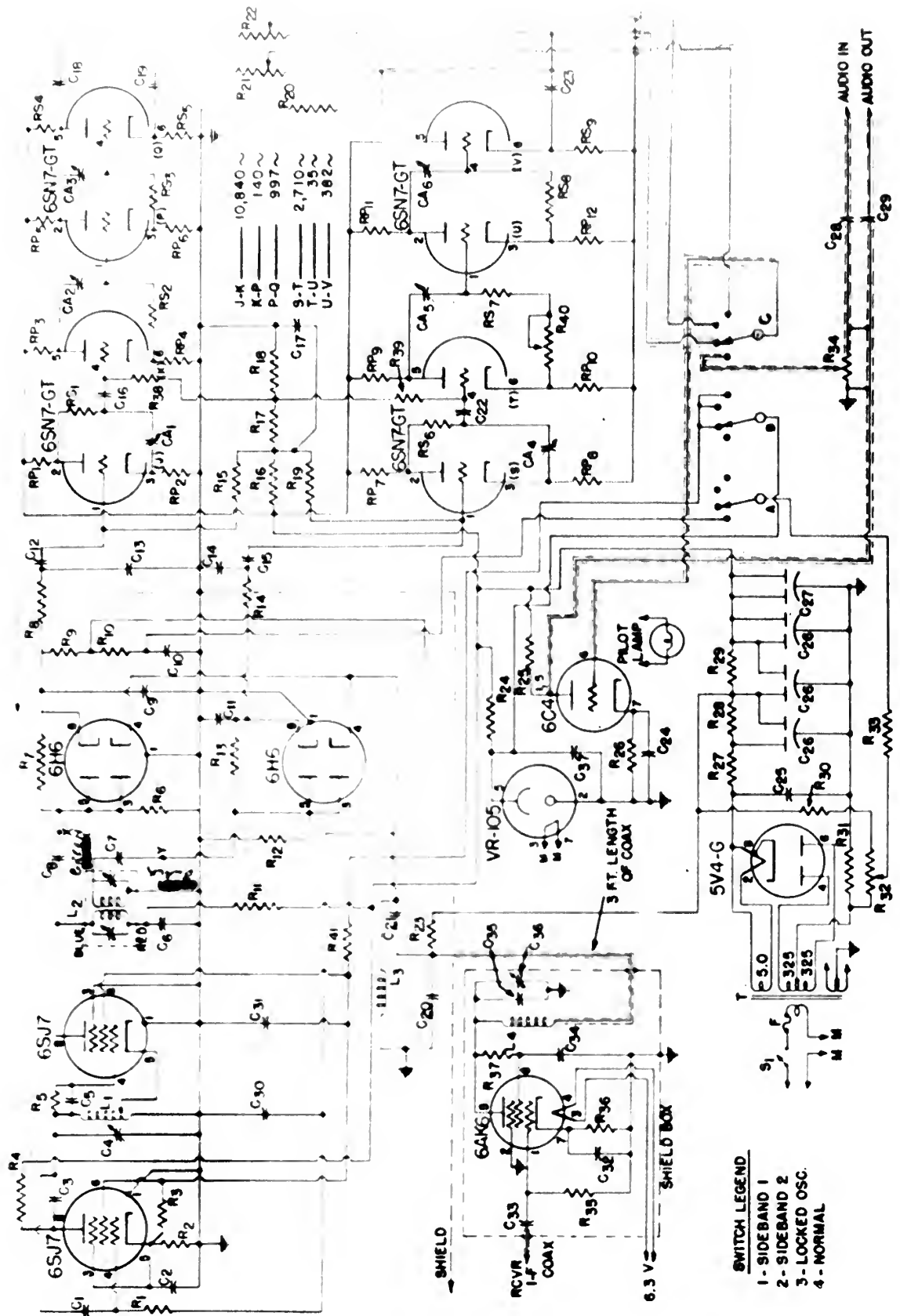


Fig. 2. Circuit Diagram of SSB Adepter





current for this tube was easily supplied by the receiver power supply.

Of the components, it should be mentioned that all resistors in the audio phase-shift networks were of the 1%, low-drift type so that accurate phase-shift could be maintained without continual alignment.

The alignment procedure was somewhat complicated. After connecting the adaptor to the receiver (a Hallicrafter S-40A was used), it was first necessary to adjust the trimmer of the final I.F. transformer to compensate for the additional capacitance introduced by the 6AK6 stage. Next, adjustment of C<sub>36</sub> gave maximum input into the adaptor. In the adaptor constructed, adjustment of L<sub>4</sub> was also necessary. The coupling of L<sub>2</sub>, which was a conventional I.F. transformer, had to be increased so that the local oscillator voltages into the two demodulator tubes were equal. Adjustment of the condensers across primary and secondary windings of L<sub>2</sub> was required to obtain accurately, the 90° phase-shift between the two demodulators. An oscilloscope, with horizontal and vertical inputs connected to the two demodulators, served as indicator for this phase aligning. When a perfect circle was obtained, the phasing was correct.



The final stage of alignment was the alignment of the two audio phase-shift networks. Again, the use of an oscilloscope served as the indicator. Each triode section of the 6SN7's with it's component parts was adjusted separately. The frequencies shown on the circuit diagram were those for which the respective elements would give  $90^\circ$  phase-shift. The adjustable elements were the indicated trimmer condensers and the adjustable resistance,  $R_{40}$ . After each element was correctly adjusted, the two systems were checked as a whole, over a frequency range of 280 cps to 6000 cps. As was mentioned before, the use of very stable resistors in the phase-shift networks is essential if this alignment procedure is not to be repeated at frequent intervals.

Results were, in general, very good. Attenuation of the undesirable sideband was 35 db below the desired one in reception of an AM transmission. SSSC reception was readily accomplished, but not with the ease that had been anticipated. Probably, the development of a certain degree of skill in the use of this adaptor is necessary to obtain optimum results. The only disappointing feature was the operation of the reactance tube in locking-in the local oscillator to the carrier frequency. A comparatively large carrier input was necessary to obtain satisfactory results.

#### Appendix VI



The local carrier would not quickly synchronize even on A.M. transmissions, if the incoming signal strength was rather weak. This was far short of the desired results of locking on a carrier 20 db below the sideband strength. It is probable that a better physical layout of the reactance tube stage to reduce stray capacitance would improve performance.

The effectiveness of SS reception of AM in reducing interference was quickly appreciated, particularly on communications channels. In several cases, signals were made understandable by use of only one sideband, where they had been entirely unintelligible, because of interference, when both sidebands were being detected.

While certain elements need additional refining, it is believed that the use of this type of an adaptor, even for reception of AM, is highly desirable for communications work. The advantages resulting from use on communications channels are not readily apparent on broadcast reception, because the types of interference are not the same for the two channels.













Thesis 13112  
Ty N37 Neman  
N. Single sideband, sup-  
pressed carrier radio  
communication.

Thesis 13112  
N37 Neman  
Single sideband, sup-  
pressed carrier radio  
communication.

06-58137

Single sideband - suppressed carrier radio



3 2768 000 98649 1

DUDLEY KNOX LIBRARY